CER-1511

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FINAL REPORT

FOR

QUARTZ CRYSTAL PHASE-STABILITY MEASURING EQUIPMENT

(R-F PHASE-STABILITY ANALYZER)

Report No. 2?
Contract: DA-36-039-sc-78310
Task Nr. 3A-99-15-004-03
Department: Electronic Components Research
Division: Solid-State and Frequency Control
Branch: Piezoelectric Crystal and Circuitry
PR and C No. 60-[illegible]1D-40004(59/[illegible]209M)
Period Covered by Report: Final Report,
1 February 196? to 19 October 1961
U.S. Army Signal Research and Development Laboratory
Fort Monmouth, New Jersey

COLLINS

COLLINS RADIO COMPANY
Cedar Rapids, Iowa

FINAL REPORT
FOR
QUARTZ CRYSTAL PHASE-STABILITY MEASURING EQUIPMENT
(R-F PHASE-STABILITY ANALYZER)

Report No. 21
Contract: DA 36-039-sc-78310
Signal Corps Technical Requirements No. SCL-6259A, Dated 20 April 1959
Task No.: 3A-99-15-004-03
Department: Electronic Components Research
Division: Solid-State and Frequency Control
Branch: Piezoelectric Crystal and Circuitry
PR and C No.: 60-ELE1D-40004(59/4809M)
Period Covered by Report: Final Report, 1 February 1960 to 19 October 1961
Object of Research: The Design and Development of Quartz Crystal Phase-Stability Measuring Equipment

Prepared By
H. P. Brower
M. E. Frerking
J. C. Schmitt

COLLINS RADIO COMPANY
Cedar Rapids, Iowa

A PUBLICATION OF
THE RESEARCH AND DEVELOPMENT LABORATORIES
COLLINS RADIO COMPANY
Cedar Rapids, Iowa

Printed in the United States of America

AD ____________ Accession No. ____________

Unclassified

Collins Radio Company, Cedar Rapids, Iowa

QUARTZ CRYSTAL PHASE-STABILITY MEASURING EQUIPMENT
H. P. Brower

Final Report, 1 February 1960 to 19 October 1961, 107 pp-Illus-Graphs (Contract DA-36-039-sc-78310) Task Nr. 3A-99-15-004-03, Unclassified Report.

Report covers development and construction of equipment to measure the phase stability of quartz crystal units in the frequency range from 1 mc to 110 mc while being subjected to 10 g vibration from 20 cps to 2000 cps. This equipment covers the entire frequency range with a unique tunable phase detector and a multiband reference oscillator in a phase-locked loop system. A crystal-vibration-test oscillator operates the crystal under vibratory test at the end of a double coaxial transmission line. Details of development problems, circuit operation, and circuit analysis are presented. The equipment has a resolution of 0.75 millidegree at 1 mc and 40 millidegrees at 110 mc. This equipment may be used not only for the development of crystal units but also for oscillator circuits, frequency synthesizers, and other r-f signal sources requiring increased phase stability.

1. Measuring Equipment Quartz Crystal Phase-Stability Design Data
2. Contract DA-36-039-sc-78310

AD ____________ Accession No. ____________

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Collins Radio Company, Cedar Rapids, Iowa

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2. Contract DA-36-039-sc-78310

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1. PURPOSE.

The purpose of this contract was to provide for the design, development, and fabrication of laboratory test equipment which will measure the phase stability of quartz crystal units during vibration. The equipment must test quartz crystals mounted in HC-6 holders, HC-18 holders, or T-5 1/2 glass bulbs, at any frequency in the range from 1 mc to 110 mc. The apparatus is to be capable of measuring peak phase deviation with a resolution which varies linearly from 3.13 millidegrees (55 microradians) at 1 mc to 344 millidegrees (6000 microradians) at 110 mc for a 10 g peak acceleration in the vibration frequency range from 20 cps to 2000 cps.

The project was divided into two phases: (1) a study phase to yield a set of design plans and (2) a construction phase consisting of the development and fabrication of three engineering models.

The design plans of phase I were to present two alternative approaches for the design of the final equipment. The first approach was to consider the use of accessory equipment to adapt the existing 32-mc fixed-frequency phase-stability measuring equipment to cover the 1-mc to 110-mc frequency range. The second approach was to propose a completely new system, capable of operating over the specified frequency range and providing the necessary resolution. The design plans were to be reviewed by USASRDL technical personnel and the approach selected used during the construction phase of the contract.

Phase II was to include the necessary circuit development, breadboard construction, and fabrication of three final engineering models which would meet the above performance requirements.

2. ABSTRACT.

This report covers a US Army Signal Research and Development Laboratory Contract for the development of equipment to measure the phase stability of quartz crystal units in the frequency range from 1 mc to 110 mc while being subjected to 10 g vibration from 20 cps to 2000 cps. The measurement resolution required varies linearly from 3.13 millidegrees at 1 mc to 344 millidegrees at 110 mc.

Inasmuch as the equipment is capable of performing measurements on any radio-frequency signal, the apparatus has been called an R-F Phase-Stability Analyzer. Basically, the system uses a tunable phase detector and a reference oscillator in a closed loop. The reference oscillator is phase locked to the average frequency of the input signal. An integrating network is employed between the output of the tunable phase detector and the reference oscillator to give the desired bandwidth to the loop.

The heart of the system is the tunable phase detector which uses several unique circuits to provide operation over the entire 1.0-mc to 110-mc frequency range. The high resolution requirements have necessitated the use of low-noise circuitry and rugged components which are relatively insensitive to microphonics.

One input signal to the tunable phase detector is supplied by the reference oscillator which is a crystal-controlled unit designed for maximum phase stability and which contains the voltage-variable reactance for the control loop. This assembly uses one 4-band oscillator to cover the 1-mc to 30-mc range and a second 4-band oscillator to cover the 30-mc to 110-mc range. A common cathode follower capable of supplying 0.5 volt for the 100-ohm input of the tunable phase detector is switched to the oscillator in use.

The second input to the tunable phase detector is supplied by the crystal-vibration-test oscillator, which is similar to the reference oscillator but has the voltage-variable reactances replaced with fixed capacitors. The crystal under test is connected to this test oscillator by means of a double coaxial cable, the length of which must be selected for the particular test frequency range. This test oscillator may be mounted near the vibration table but remote to the equipment cabinet.

The output of the phase detector is connected through a band-limiting audio amplifier to an oscilloscope, which is the principal readout device. The oscilloscope is calibrated in terms of degrees of phase deviation per volt by opening the system control loop and observing the peak-to-peak output voltage of the tunable phase detector.

Details of development problems, explanation of circuit operation, and circuit analysis are presented for the tunable phase detector, reference oscillator, crystal-vibration-test oscillator, and the complete phase-locked loop system. Pertinent information also is given on other parts of the system, such as the audio amplifier, oscilloscope, power supplies, and cabinet. System operation is detailed, and performance data are presented.

The system performance is at least three times better than that required by the technical specifications. Phase-deviation resolution of 0.75 millidegree at 1 mc and 40 millidegrees at 110 mc was measured in the 4-kc system bandwidth.

Application of this equipment will facilitate the development of phase-stable crystal units. It will be a very useful measurement tool for the development of circuits and components for crystal oscillators, frequency synthesizers, and other radio-frequency signal sources demanding increased phase stability.

3. PUBLICATIONS, LECTURES, REPORTS, AND CONFERENCES.

3.1 Reports. During phase I of this contract, seven monthly letter reports (Collins Interim Development Report IDR-532-1 through IDR-532-7) were written. The seventh report (IDR-532-7) constituted the design plans for the contract and contained the six previous monthly letter reports as an appendix.

The theoretical work for the study phase was subcontracted to Newell Engineering Company, Cedar Rapids, Iowa. This work was a coordinated effort between Dr. Newell, and his associates, and H. P. Brower of Collins Radio Company. The results of these efforts, together with various experimental work and system studies conducted by Collins, were presented in the monthly reports and the design plans.

Monthly letter reports (IDR-532-8 through IDR-532-20) also have been written to cover the construction phase during which the development and fabrication of the final engineering models were completed.

3.2 Conferences.

a. Date: 17 March 1960

Place: USASRDL, Fort Monmouth, New Jersey

Personnel in Attendance:

Owen P. Layden Stanley S. Schodowski Marvin Bernstein Dennis Pochmerski George H. Gougoulis Lewis S. Nelson	USASRDL, Frequency Control Division
William Felty	Collins Radio Company Red Bank, New Jersey
Darrell E. Newell, Consultant	State University of Iowa Iowa City, Iowa
H. P. Brower	Collins Radio Company Cedar Rapids, Iowa

Purpose of Meeting: To discuss the details of the study phase of this project.

Conference Details: The general program to be followed in the study phase of this contract was outlined. Particular attention was given to the work to be subcontracted to and conducted by the Newell Engineering Company under the direction of Dr. D. E. Newell. An outline for this work was presented in the letter report for February 1960 (IDR-532-1).

In discussing the several systems presented in the original Collins proposal for this contract, it was agreed that the use of an offset crystal should be avoided and would be used only as a last resort. It was requested that, if possible, an output be provided on the future equipment so that a frequency measurement on the quartz crystal under test also could be made. It also was agreed that this would be investigated when a thorough study of systems was undertaken later in the study phase.

The existing Motorola phase-stability measuring equipment was inspected, and the performance was observed. Considerable discussion centered around the problem of transporting one of these measuring systems from Kansas City, Kansas, (Midland Manufacturing Company) to Cedar Rapids, Iowa.

It was suggested by USASRDL personnel and agreeable to Collins personnel that another conference be held when the study phase is completed but prior to writing the final study phase report. In this connection, it was further agreed that the monthly letter reports would be more voluminous and complete than usually required. This would make it possible to have the majority of the study phase work recorded by the time the study phase report was begun. As a result, the study phase report (design plans) will contain a discussion of the work, the conclusions, and the previous letter reports in the appendix.

b. Date: 18 May 1960

Place: Midland Manufacturing Company
Kansas City, Kansas

Personnel in Attendance:

George Gougoulis	USASRDL
Lewis Nelson	USASRDL
H. P. Brower	Collins Radio Company

Purpose of Meeting: To obtain one rack of 32-mc crystal phase-stability measuring apparatus.

Results: One rack of the existing 32-mc crystal phase-stability measuring apparatus was obtained from the Midland Manufacturing Company, Kansas City, Kansas, and transported to Cedar Rapids in a Collins Radio Company aircraft. Mr. H. P. Brower, Collins Radio Company, met Mr. George Gougoulis and Mr. Lewis Nelson from USASRDL at Midland Manufacturing Company and observed the operation of the rack. Mr. Gougoulis and Mr. Brower accompanied the rack shipment to Cedar Rapids.

c. Date: 19 May 1960

Place: Collins Radio Company
Cedar Rapids, Iowa

Personnel in Attendance:

George Gougoulis	USASRDL
D. E. Newell	Newell Engineering Company Cedar Rapids, Iowa
H. P. Brower	Collins Radio Company

Purpose of Meeting: To reassemble equipment and review study phase progress.

Results: Mr. Gougoulis and Mr. Brower reassembled the rack of 32-mc crystal phase-stability measuring equipment and checked it for proper operation. Mr. Gougoulis discussed various aspects of the project with Dr. Newell and Mr. Brower. The work conducted on the study phase to date was reviewed.

d. Date: 6 and 7 July 1960

Place: USASRDL, Fort Monmouth, New Jersey

Personnel in Attendance:

Owen P. Layden	USASRDL, Frequency Control Division
Dennis Pochmerski	
George H. Gougoulis	
Lewis S. Nelson	
Marvin Bernstein	
Stanley S. Schodowski	
Darrell E. Newell	Collins Radio Company
H. Paul Brower	Collins Radio Company

Purpose of Meeting: To discuss various systems which had been proposed for the final apparatus.

Results: Various proposed systems were discussed. It was decided that all of these would be reported in the study phase reports.

The problem of noise in crystal oscillators was discussed with Dr. Erich Hafner, and a copy of the paper he presented on this subject at the Fourteenth Annual Frequency Control Symposium was obtained.

e. Date 21 October 1960

Place: Collins Radio Company
Cedar Rapids, Iowa

Personnel in Attendance:

Owen P. Layden	USASRDL
S. M. Morrison	Collins Radio Company
A. E. Anderson	
N. K. Garnatz	
H. P. Brower	
D. E. Newell	Newell Engineering Company Iowa City, Iowa

Purpose of Conference: To discuss the project design plans which had been submitted to USASRDL as a result of the study phase of the project.

Results: The design plans were discussed in detail. An artist's sketch of the proposed ultimate equipment design was presented and used as a basis for discussing the operation of the apparatus. All persons present agreed that the ultimate performance of this equipment depended on the satisfactory development of a tunable phase detector.

IDR-532-7 indicated a need for clarification of the request for an output jack for frequency measurement. After discussion of this problem, it was agreed that the output jack should be provided at the output of the cathode follower. This would allow connection of an external filter with a different pass band than that provided within the audio filter-amplifier circuit.

It was agreed to provide output jacks for external measurement of the frequency of both the reference and crystal-vibration-test oscillators.

The crystal-vibration-test oscillator was discussed, and it was agreed that it may be necessary to remove the oscillator circuitry from the vibration table if it is proven that vibration of the circuitry is affecting the test data adversely.

Mr. Layden advised that official approval of the design plans for construction of a completely new measurement system using a tunable phase detector would be received through regular channels after his return to Fort Monmouth.

Mr. Garnatz (Collins Radio Company Contract Administrator) discussed with Mr. Layden the problem of the contract period for the construction phase. It was agreed that the answer to this contract problem would be handled through regular channels after receipt of the approval of the design plans.

f. Date: 5 April 1961

Place: Collins Radio Company
Cedar Rapids, Iowa

Personnel in Attendance:

George Gougoulis	USASRDL
S. M. Morrison	Collins Radio Company
A. E. Anderson	
H. P. Brower	
M. E. Frerking	
D. L. Linman	

Purpose of Conference: To show Mr. Gougoulis the progress on the project to date.

Results: Due to the difficulties with the JFD Electronics Corporation tubular glass capacitors, we were not in the best position to demonstrate adequately the system performance. However, we were able to demonstrate compliance with the system resolution requirement of .003 degree peak phase deviation at 1 mc.

Mr. Gougoulis expressed concern over the possibility of holes existing in the performance frequency range of the reference oscillator. Up to that time a problem had not been observed in that unit or the phase detector.

Mr. Gougoulis was concerned about accessability to the phase detector assembly and made several constructive criticisms regarding its mechanical design. Appropriate changes in the design of the engineering model of the phase detector have been incorporated.

g. Date: 19 May 1961

Place: USASRDL, Fort Monmouth, New Jersey

Personnel in Attendance:

O. P. Layden	USASRDL, Frequency Control Division
G. Gougoulis	
D. Pochmerski	
R. W. Britton	Collins Radio Company
A. E. Anderson	
H. P. Brower	

Purpose of Conference: To discuss the technical progress on the project to date and the expected over-expenditure of contract funds.

Results: Data showing satisfactory performance of the various system subassemblies were presented and remaining problems were discussed. Photographs of the various subassemblies in the engineering model of the R-F Phase Stability Analyzer also were shown.

h. Date: 4 and 5 October 1961

Place: Collins Radio Company
Cedar Rapids, Iowa

Personnel in Attendance:

G. Gougoulis	USASRDL
D. Pochmerski	USASRDL
A. E. Anderson	Collins Radio Company
N. K. Garnatz	
H. P. Brower	
M. E. Frerking	
J. C. Schmitt	
H. Rice	

Purpose of Conference: To demonstrate construction and operation of the equipment and compliance with the contract technical requirements as specified in SCL-6259A.

Results: Physical inspection of the design and construction of the equipment was conducted. Application of the equipment then was made in the Collins Environmental Area. Crystals at various frequencies were mounted on the vibration table and their phase deviation was observed.

One 32-mc unit brought by Mr. Gougoulis was tested and the results compared within 25 percent of the values obtained on the fixed frequency phase stability measurement equipment at USASRDL. Resolution of the equipment was demonstrated at 1 mc, 32 mc, and 110 mc and was found to be considerably better than that required by the technical specifications.

Mr. Gougoulis and Mr. Pochmerski worked with the equipment to become familiar with the tuning and calibration procedures. During this period, Mr. Gougoulis suggested an alternate calibration technique, which is included in this report and in the instruction manual.

Mr. Gougoulis and Mr. Garnatz, Collins contract administrator, discussed several questions raised by USASRDL regarding contract billing. These questions were answered satisfactorily.

A discussion also was conducted with Mr. Garnatz and Mr. Rice concerning the interpretation of SCL-2101K, paragraph 3.2.3.3. It was agreed that Collins would include with the final report one set of transparencies of all the model shop drawings for this equipment. This would allow USASRDL to reproduce any drawings they desired.

Mr. Garnatz advised that the USASRDL contracting officer was expecting to receive the final report in printed form about 15 October 1961. However, paragraph 4 of SCL-2101K specifies submission of a draft of the final report for approval by USASRDL before final printing. Some question remained as to whether this paragraph was applicable to this contract and Mr. Garnatz advised that he would check the problem with the USASRDL contracting officer.

4. FACTUAL DATA.

4.1 Introduction. The contract title for this project, as indicated on the cover, is Quartz Crystal Phase-Stability Measuring Equipment. However, since the apparatus is applicable for the measurement of the phase stability of any radio-frequency signal having the required amplitude, the final engineering models of the equipment carry the title R-F Phase-Stability Analyzer.

As a result of the study phase, the design plans submitted to USASRDL recommended that a complete system using a tunable phase detector be developed and constructed during phase II. After receipt of approval by USASRDL of these design plans, the phase II work proceeded.

During the course of the development, several changes in the basic system concept were made, and as a result, the block diagram of figure 1 was used in the final equipment.

This system functions in the following manner. The crystal to be tested for phase stability under vibration is mounted in a suitable holder on the vibration table and connected to the crystal vibration test oscillator by means of an appropriate length of low-noise coaxial cable. The test oscillator is adjusted for proper output when connected to one input to the tunable phase detector. Another crystal of identical frequency is plugged into the reference oscillator, which is permanently connected to the second phase detector input. After the reference oscillator is adjusted properly, the phase detector is tuned to give maximum unlocked output voltage on the oscilloscope. The loop containing the integrating network then is closed and the reference oscillator locked to the average frequency of the test oscillator. Once the loop is locked, the phase deviation of the test oscillator signal can be observed by increasing the sensitivity of the oscilloscope.

In the operation of this loop, the output signal of the phase detector contains two components. The d-c component is averaged by the integrating network and used to control the reference oscillator. The a-c component of the output is connected to an audio amplifier containing a low-pass filter. The response of the audio circuits thereby is limited to a frequency range of 20 cps to 4 kc. The output of the audio amplifier is connected to the oscilloscope which displays the phase changes in the signal under test.

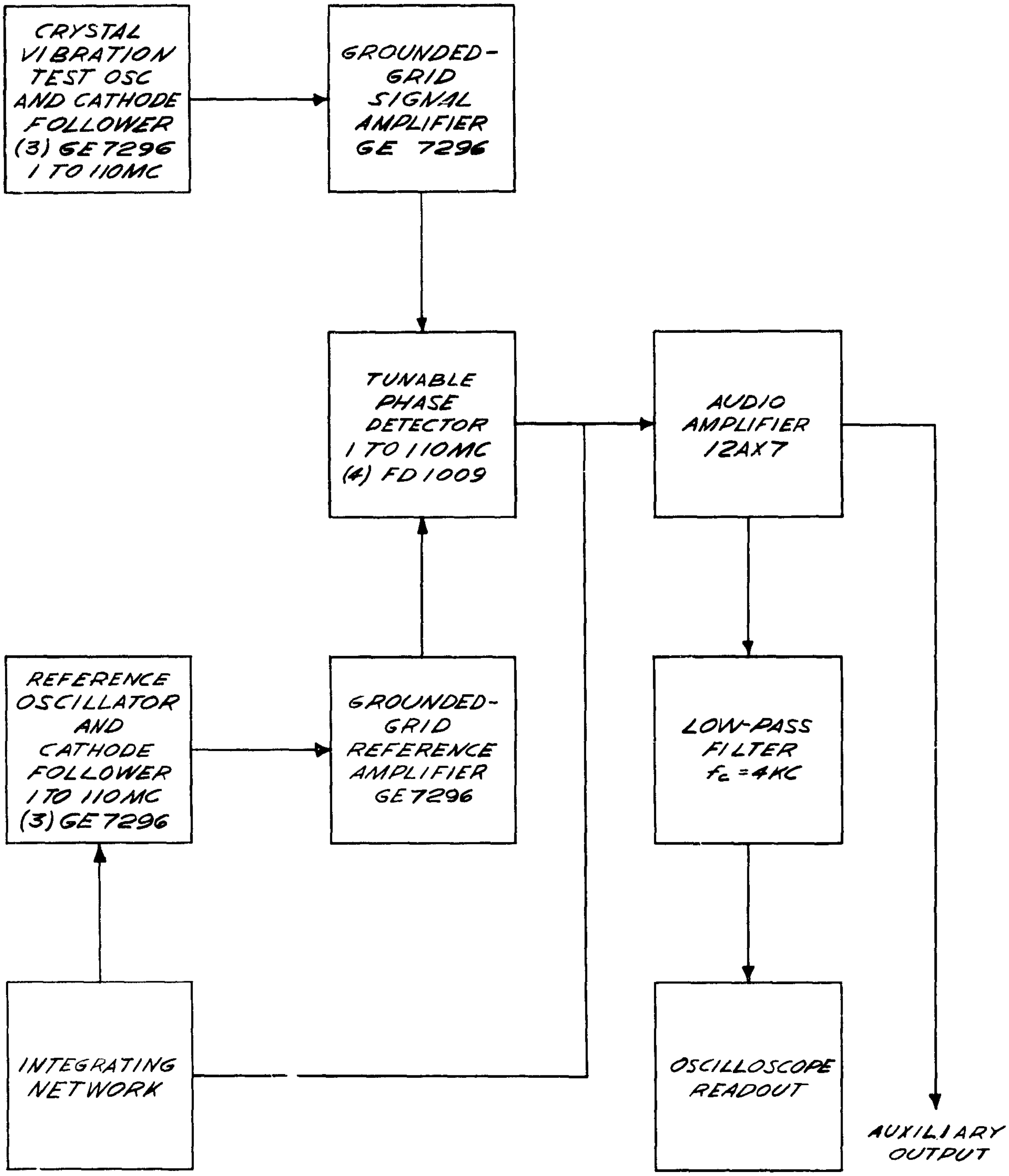


Figure 1. R-F Phase-Stability Analyzer, Block Diagram

As indicated in figure 1, an auxiliary output from the audio amplifier circuit is provided for a differentiator circuit. By means of external circuitry, changes in the frequency of oscillation of the crystal under test may be observed.

The measurement system has been assembled in a caster-mounted cabinet as shown in figures 2 and 3. The main chassis, which contains the reference oscillator, tunable phase detector, audio amplifier, and integrating network, is mounted below the oscilloscope on the cabinet turret. Below the turret is a drawer in which the crystal-vibration-test oscillator and accessories are stored. The power supply chassis is mounted at the rear of the cabinet above the drawer, as shown in figure 3. As shown, hangers are provided inside the rear door for the power and coaxial cable for the test oscillator. These cables are sufficiently long to allow the test oscillator to be placed close to the vibration table but remote to the system cabinet.

4.2 System Basic Units. As shown in figure 1 and previously discussed, the R-F Phase-Stability Analyzer is composed of the following basic units or subassemblies:

Tunable Phase Detector

Reference Oscillator

Crystal-Vibration-Test Oscillator

Audio Amplifier

Integrating Network

Power Supplies

Oscilloscope

Cabinet

The tunable phase detector, reference oscillator, audio amplifier, and integrating network are installed in the main chassis. Various views of the main chassis are shown in figures 4, 5, 6, and 7.

In the following material, a discussion of each of these units will be presented. This discussion will include the theoretical (where applicable) and practical aspects of the unit. Also, various problems and changes in design that were encountered during phase II will be presented.

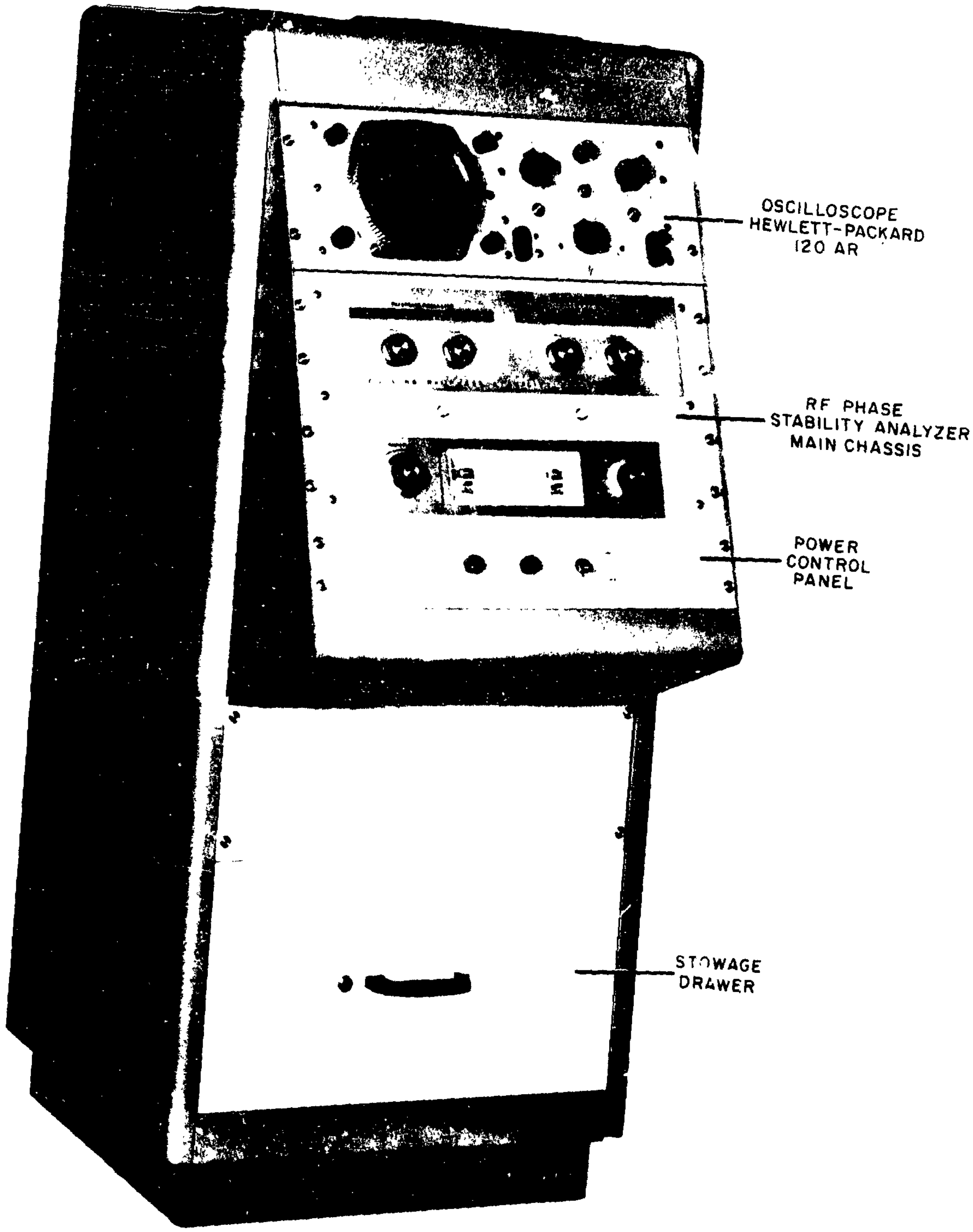


Figure 2. R-F Phase-Stability Analyzer, Front View

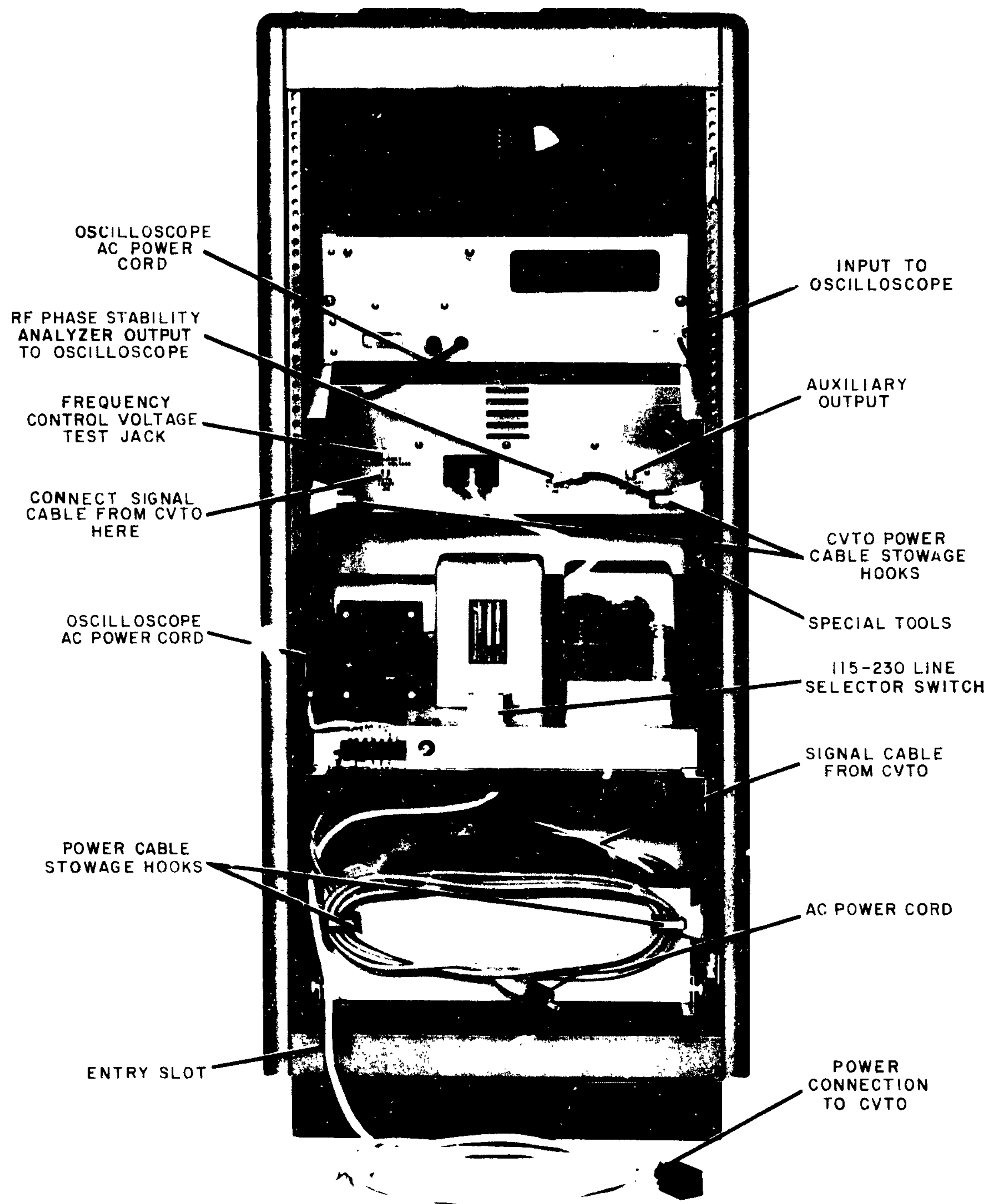


Figure 3. R-F Phase-Stability Analy ·r, Rear View

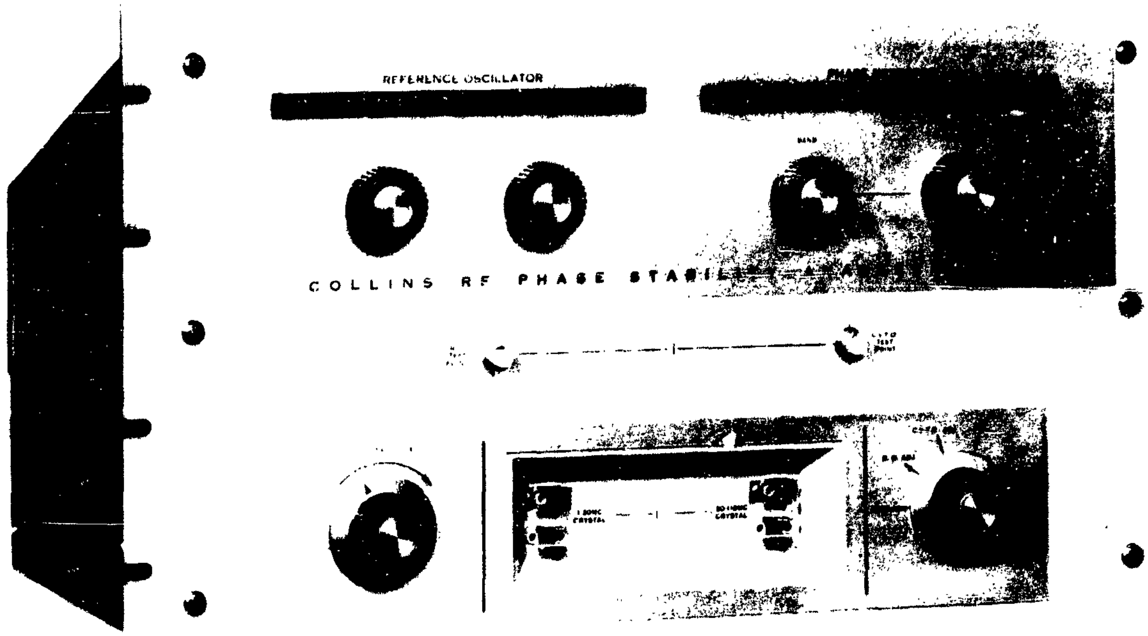


Figure 4. Main Chassis, Front Panel

4.2.1 Tunable Phase Detector.

4.2.1.1 General Discussion. The tunable phase detector is the heart of this system and functions basically as a fixed-tuned phase detector at any frequency in the range from 1 mc to 110 mc. The feasibility of a tunable phase detector, and in particular one covering such a wide frequency range, was open to serious question (see paragraph 3.2e) when proposed in the design plans since no similar circuitry was known. Tests conducted early in phase II established its workability but left a number of problems to be answered.

Several effects, which can be adequately compensated in a fixed-tuned phase detector, become significant in a tunable unit. The most important of these are the transformer primary to secondary capacity and stray capacity. Figure 8 is a basic schematic diagram of a fixed-tuned phase detector where C_s represents circuit stray capacity and C_{ps} represents primary to secondary capacity of the transformers.

Since C42 and C56 are relatively large, the primary to secondary capacity (C_{ps}) of the center-tapped signal transformer is effectively in parallel with C51, the tuning

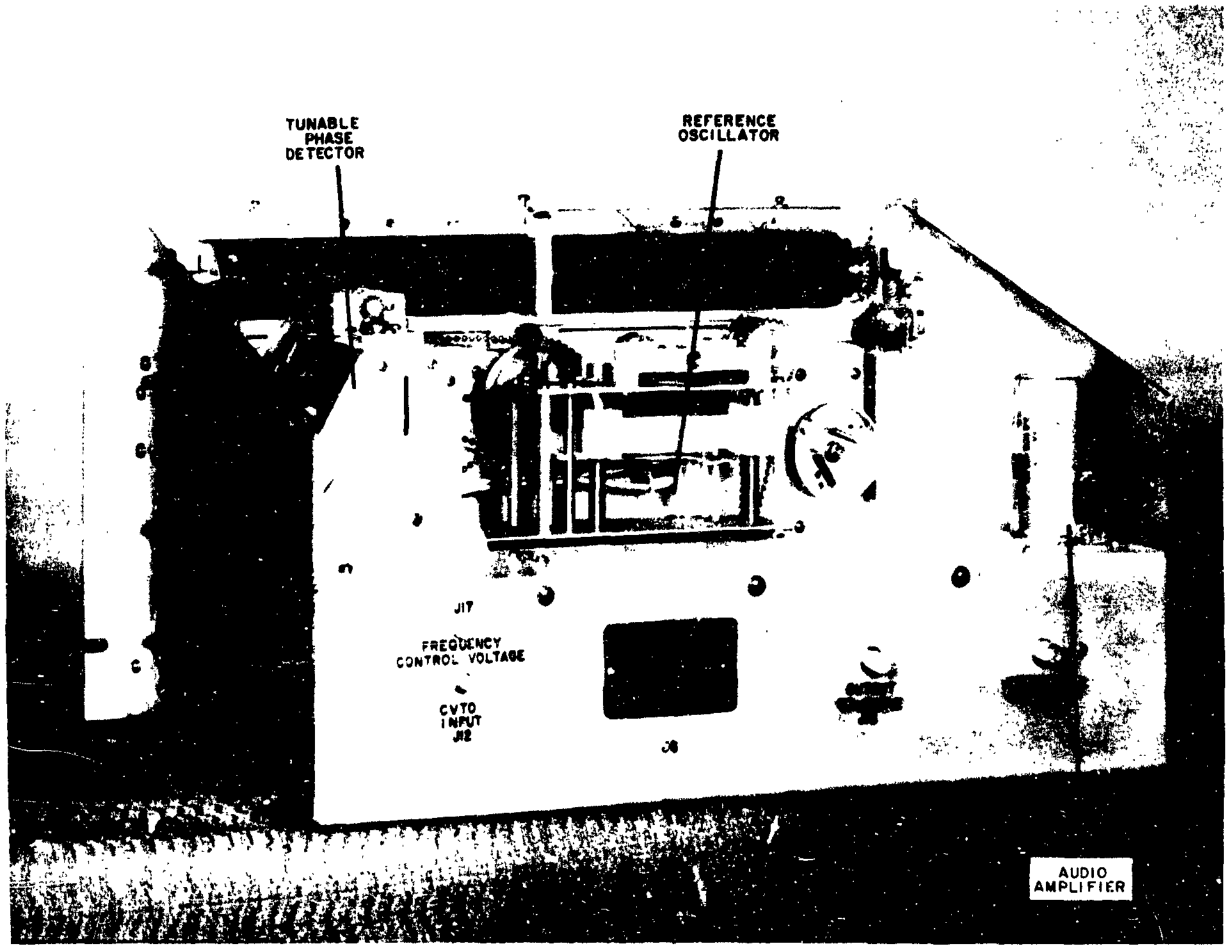


Figure 5. Main Chassis, Rear View

capacitor for the reference transformer. During early development work, this capacity combined with stray capacity was measured to be 40 pf at 1 mc. In a tunable phase detector this capacity is undesirable, since it seriously restricts the frequency band over which a given coil may be tuned. Therefore, to correct this situation, the circuit of figure 9 is used in a tunable phase detector. The use of the two r-f chokes (L21 and L22) reduces the shunting effect on the reference transformer to approximately 10 pf. These chokes (2 millihenries) do not adversely affect the flow of either the d-c or a-f currents. These two r-f chokes must be chosen to provide a large impedance over the entire operating range of the phase detector. The first choice for these inductors was to use a 5-mh and a 500-uh inductor in

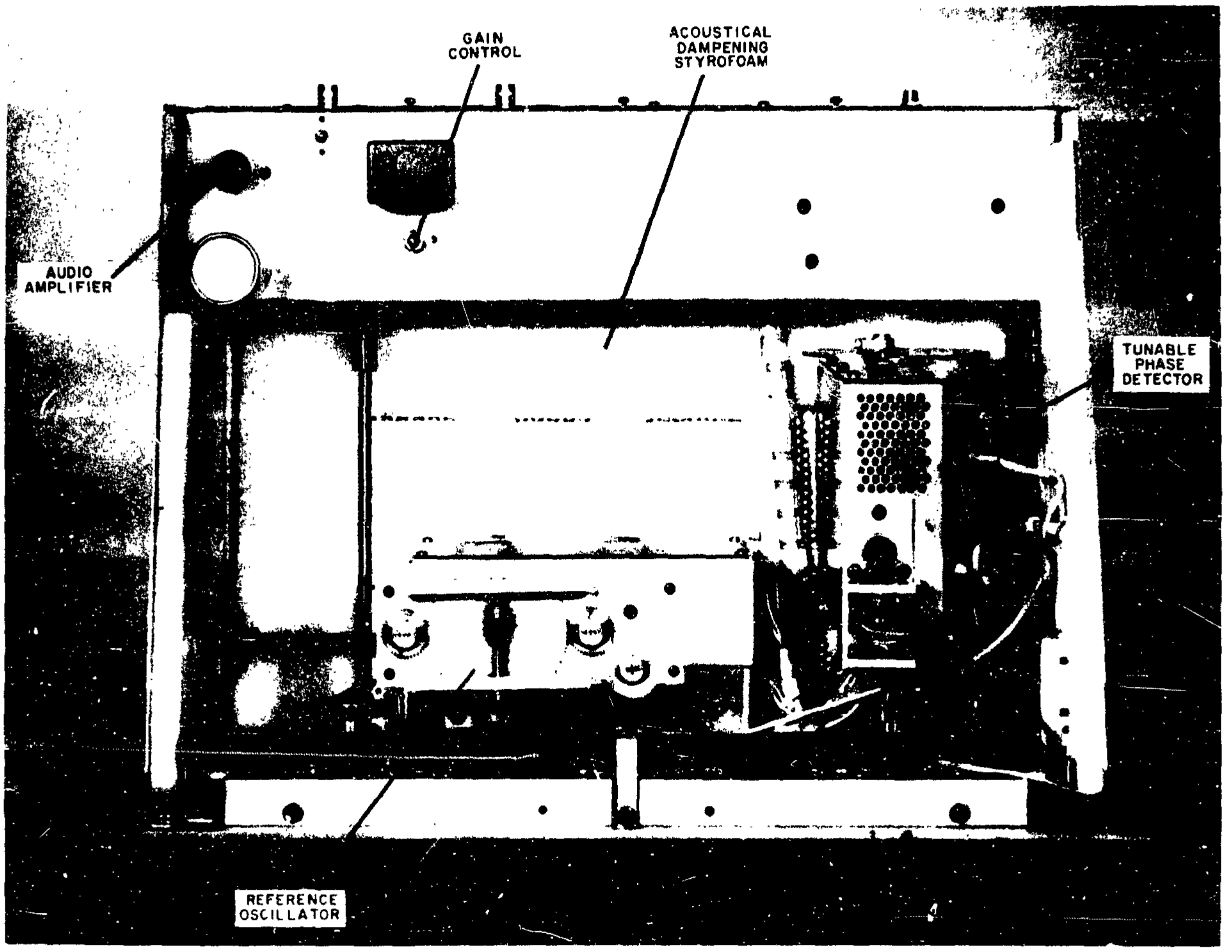


Figure 6. Main Chassis, Top View

series for each path. Because of the distributed capacity and the resulting parallel resonant frequency of each choke, the 5-mh inductor was to provide a high impedance in the low-frequency region and the 500-uh inductor was to provide a high impedance in the high-frequency region. Unfortunately, the combination was found to become series resonant at approximately 3 mc, greatly reducing the phase detector output voltage at that frequency. A single 2-mh inductor was found to perform satisfactorily over the frequency range.

At frequencies above approximately 6 mc, another effect of the stray capacity manifests itself. Referring again to figure 8, it can be seen that an r-f path exists from the primary of the signal transformer through the reference transformer to ground by way of C_{ps} of the signal transformer and C_s. This circuit couples power from the signal

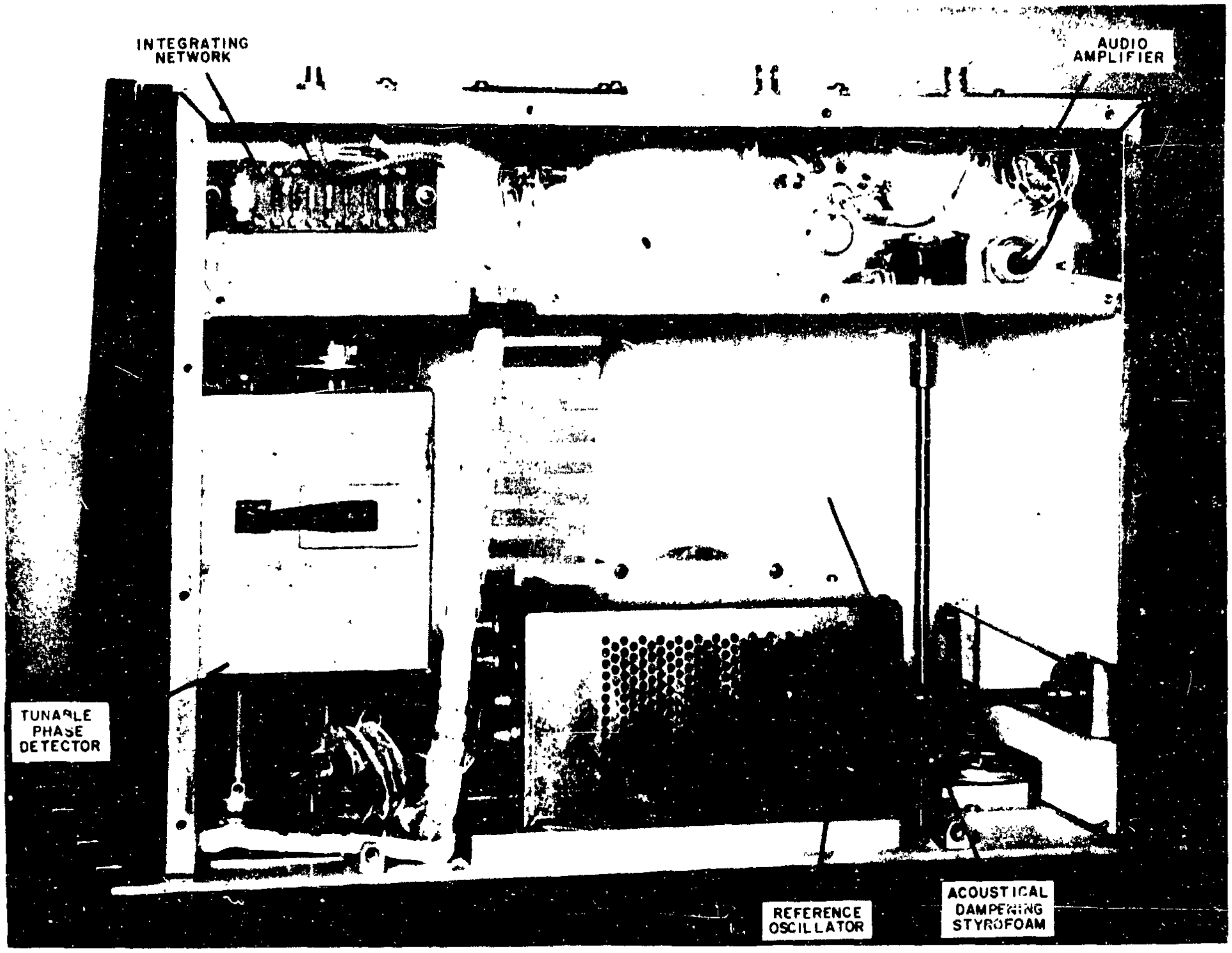


Figure 7. Main Chassis, Bottom View

transformer to the reference transformer. Since the voltage so induced in the reference transformer is not 90 degrees out of phase with the signal voltage, an output results from the phase detector in the form of an unbalance. This undesirable effect is corrected by the use of C52 and C53 as shown in figure 10. In this schematic, the desirable r-f path is shown as heavier lines. The r-f connection between the reference and the signal transformers is completed by two 47-pf capacitors (C52, C53) in series. This arrangement eliminates the stray r-f path between the transformers, thus correcting the unbalance. The d-c and a-f path is maintained by shunting the capacitors with the 500 microhenry chokes, L19 and L20.

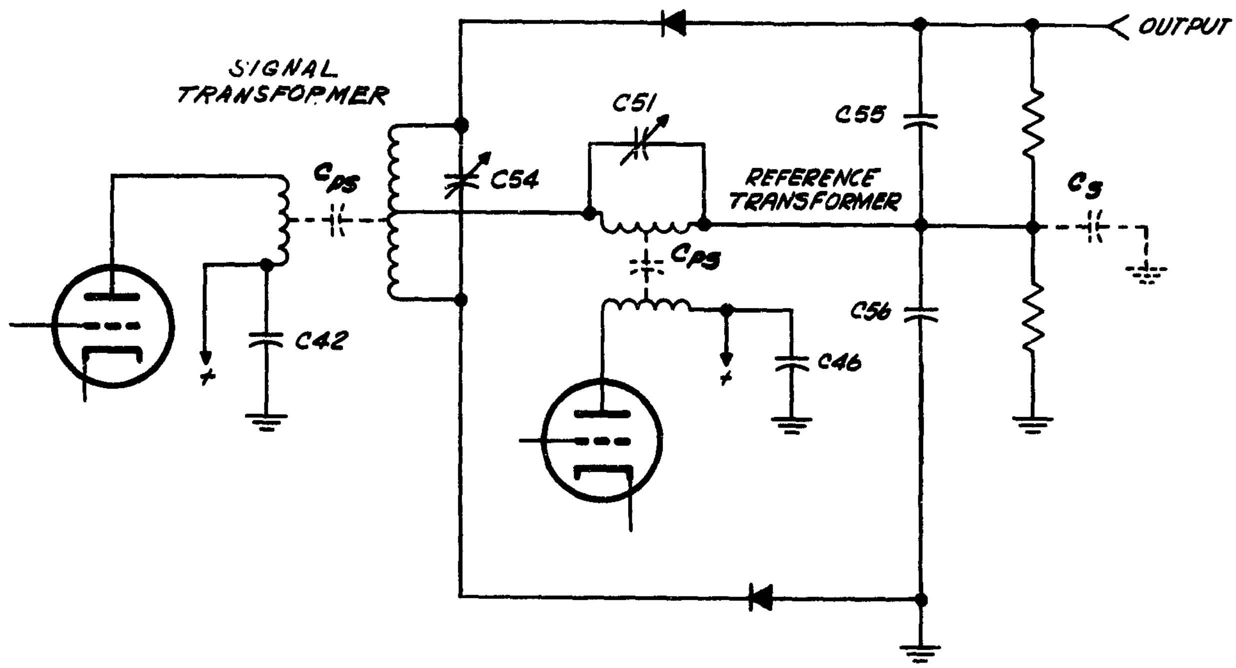


Figure 8. Basic Fixed Tuned Phase Detector

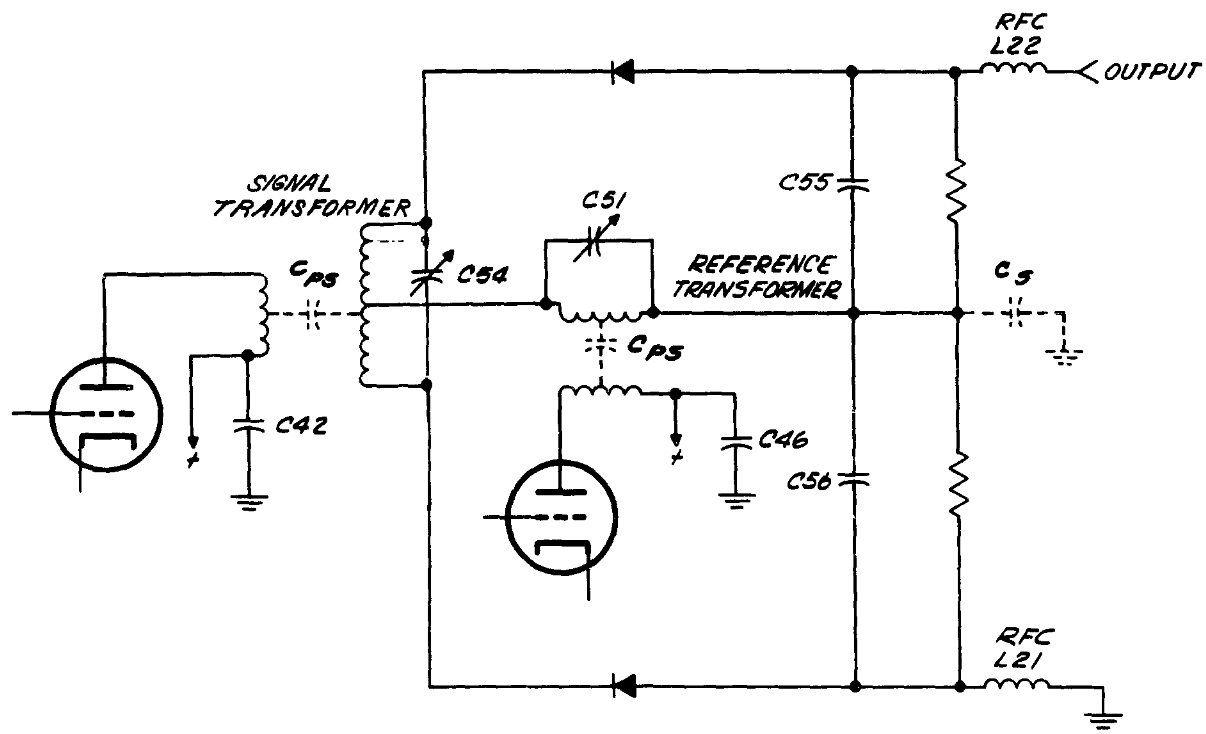


Figure 9. Tunable Phase Detector Circuit to Reduce Effects of C_{ps}

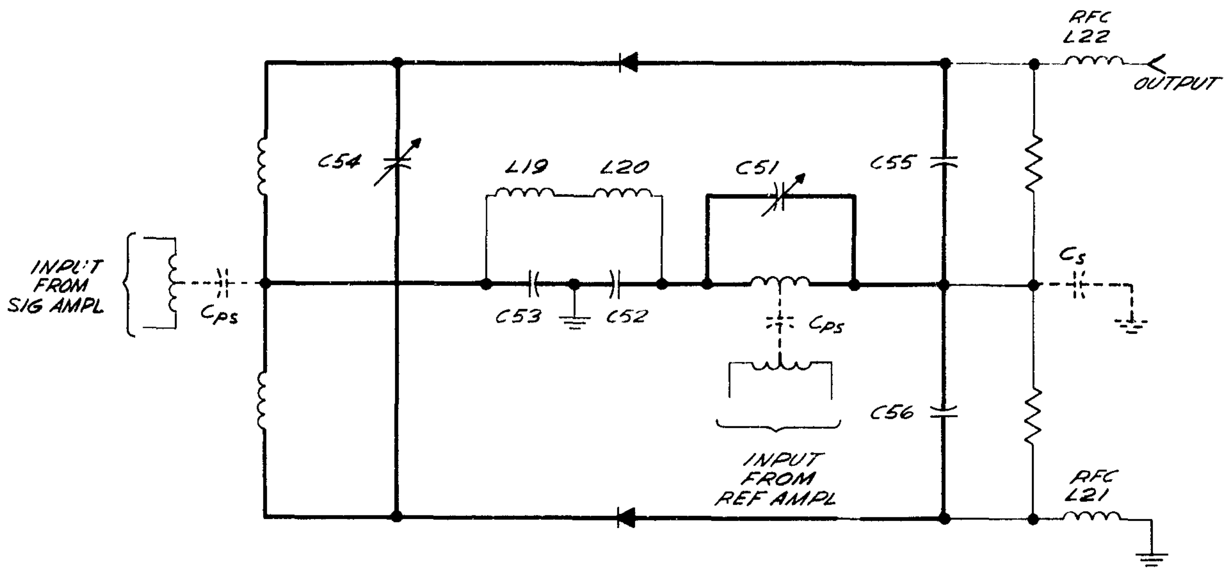


Figure 10. Tunable Phase Detector Circuit for Reducing the Unbalance Due to Effects of C_{ps} and C_s

The phase detector transformers were manufactured to Collins specifications by Communications Coil Company, Chicago, Illinois. The secondary was designed to tune the specified frequency range with a total capacitance of 25 to 180 picofarads and a load impedance of 100K ohms. The primary impedance is 9K ohms with a coupling coefficient of 0.6 or larger. Each transformer has an adjustable core for balancing the output voltage. The windings were designed to mount into the TV turret strips and have the shortest possible leads to the terminals.

At frequencies above 30 mc, difficulty was encountered in constructing phase detector transformers possessing the desired characteristics. It was found that the coefficient of coupling between the primary and secondary could not be made large enough to obtain a sufficient secondary voltage. It also was found that the resonant frequency of the primary windings influenced the tuning of the transformer secondaries upsetting tracking of the two circuits. As a result, the phase detector of figure 11 was developed for use in the vhf region. This circuit uses shunt-fed tank circuits in place of the transformers. From this schematic, it can be seen that the desirable r-f path has not been changed and the phase

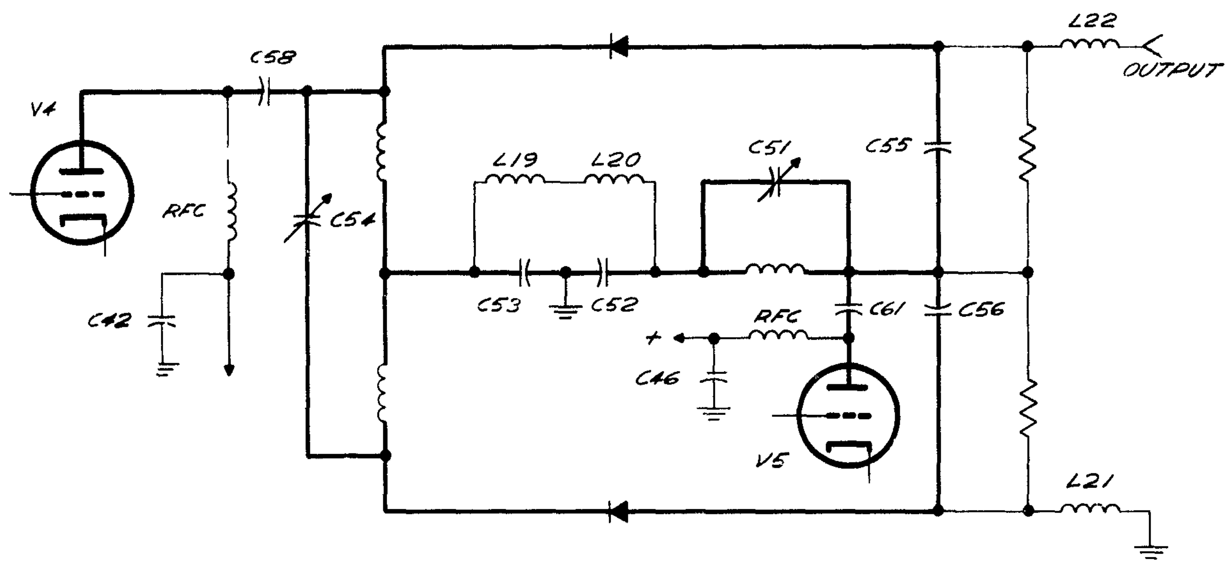


Figure 11. VHF Tunable Phase Detector Circuit

detector functions basically the same as at the lower frequencies. A complete schematic diagram of the 1-mc to 110-mc tunable phase detector, incorporating all the circuits described, is shown in figure 12.

In calibrating the complete tunable phase detector with the transformers and the coils for all the bands mounted in the turret, several undesirable holes were found in the output voltage. These holes were due to natural resonant frequency of the transformer or coil assembly on the turret insert adjacent to the strip in use. Altering the size of the padder capacitors on the appropriate turret strips and readjusting the main tuning capacitors moved these holes outside the tuning range in use. Any future design of a tunable phase detector should include either shields between the coils or a method of shorting the coil assemblies not in use.

4.2.1.2 Mechanical Design. The mechanical construction of the complete tunable phase detector assembly is shown in figure 13. Switching of the signal and reference transformers is accomplished by means of the TV turret shown in figure 14. This turret is mounted directly behind the printed circuit board in order to reduce lead length and preserve circuit

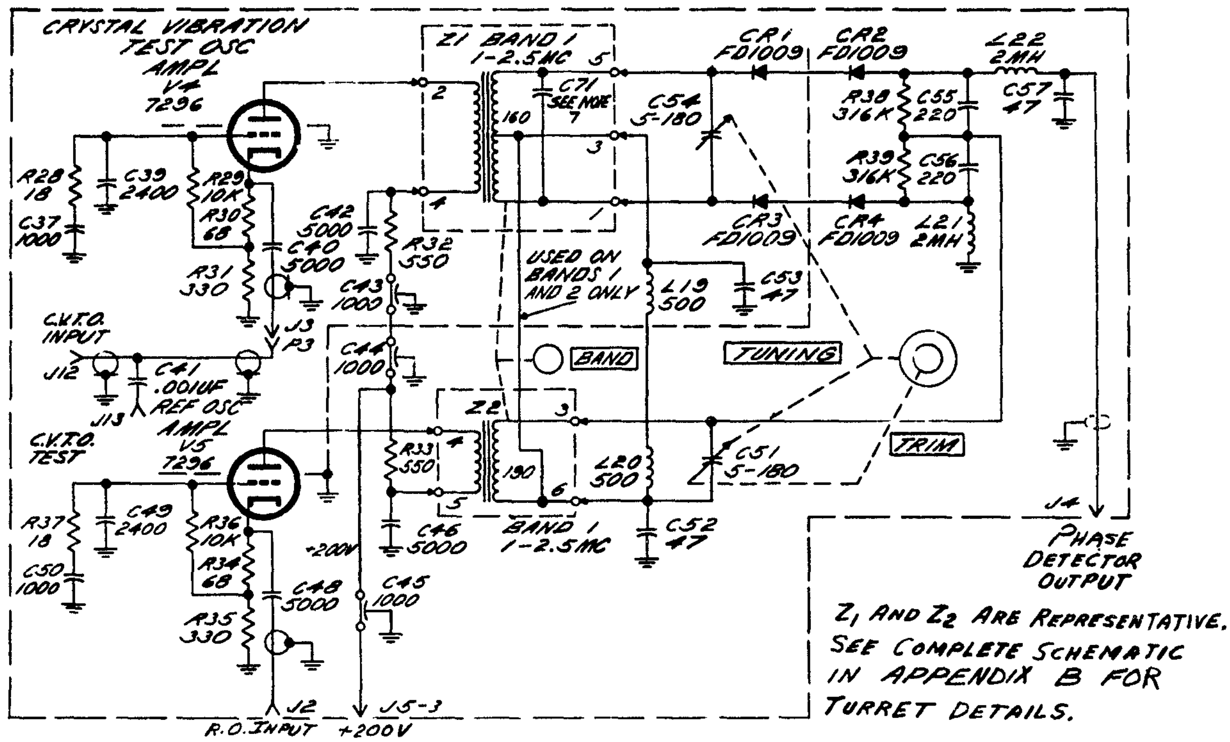


Figure 12. 1-Mc to 110-Mc Tunable Phase Detector, Schematic Diagram

balance and symmetry. Note the complete shielding and the construction provided for the grounded-grid circuitry.

In addition to the unique circuitry used to eliminate the above-mentioned problems, a great deal of care in component selection was dictated by the extreme phase resolution requirements of the specifications. As outlined in the study phase report, it is necessary to use triode vacuum tubes to achieve the low noise level required.

It was determined that extremely rugged tubes are required to prevent microphonic pickup from phase-modulating the signal. This problem was eliminated by the use of ceramic tubes. As indicated in the study phase report, it was intended originally to use GE 7077 ceramic triodes. However, further development indicated that a larger plate dissipation would be required to increase the phase detector output. As a result, GE 7296 ceramic triodes were used in the final equipment. The 7296 maintains its mutual conductance

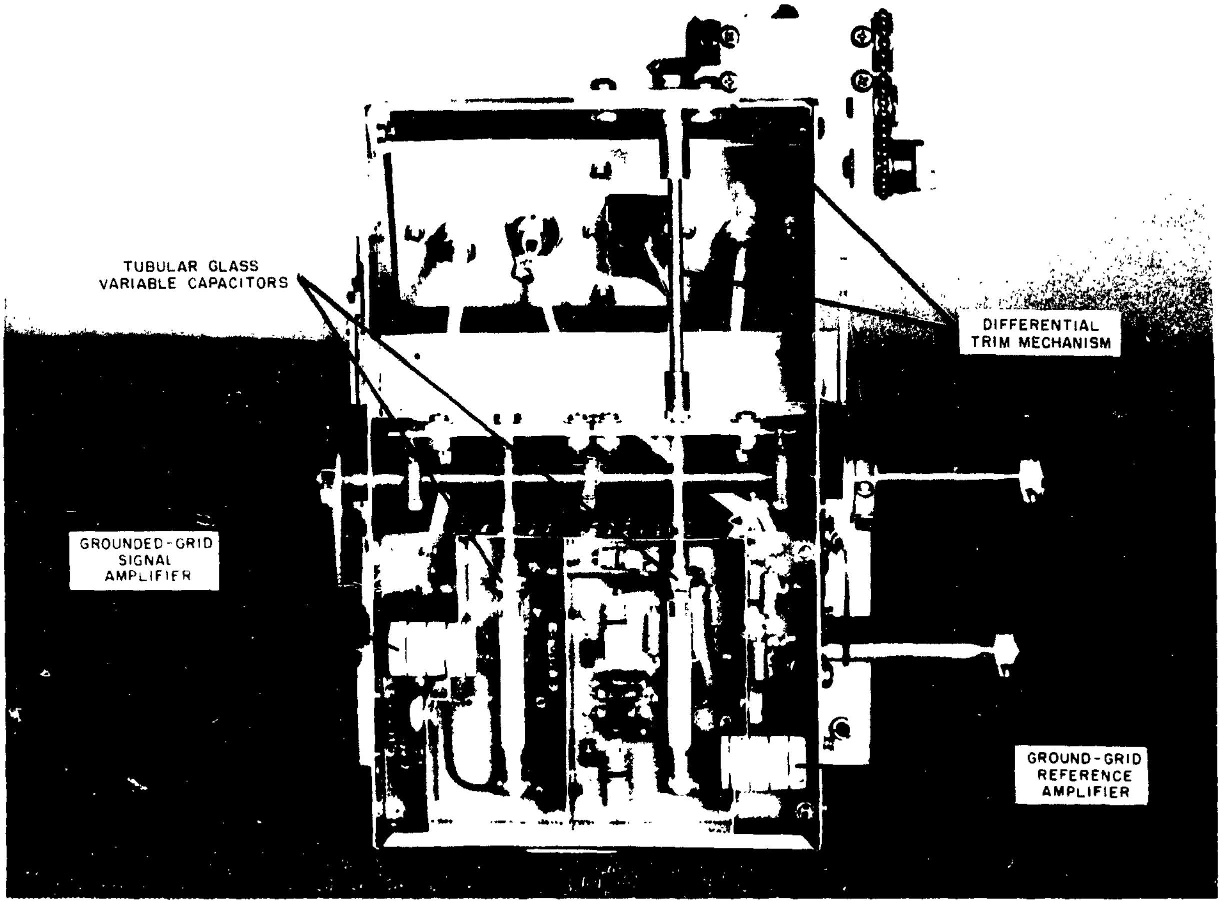


Figure 13. Tunable Phase Detector Assembly, Front View

of 15,000 micromhos well into the vhf region. As a result, a considerable amount of difficulty was experienced with parasitic oscillations in the 200- to 400-mc range. Since conventional parasitic suppressors were found to degrade performance of the unit around 100 mc, such techniques could not be used. As a result, the physical layout of the phase detector board became extremely important. These considerations, combined with the grid grounding network (shown in figure 12), eliminated the parasitic oscillations.

Considerable difficulty also was encountered in the selection of a tuning capacitor. It was found that air variable capacitors could not be used because of susceptibility of the capacitor plates to microphonics and mechanical vibration. The use of variable

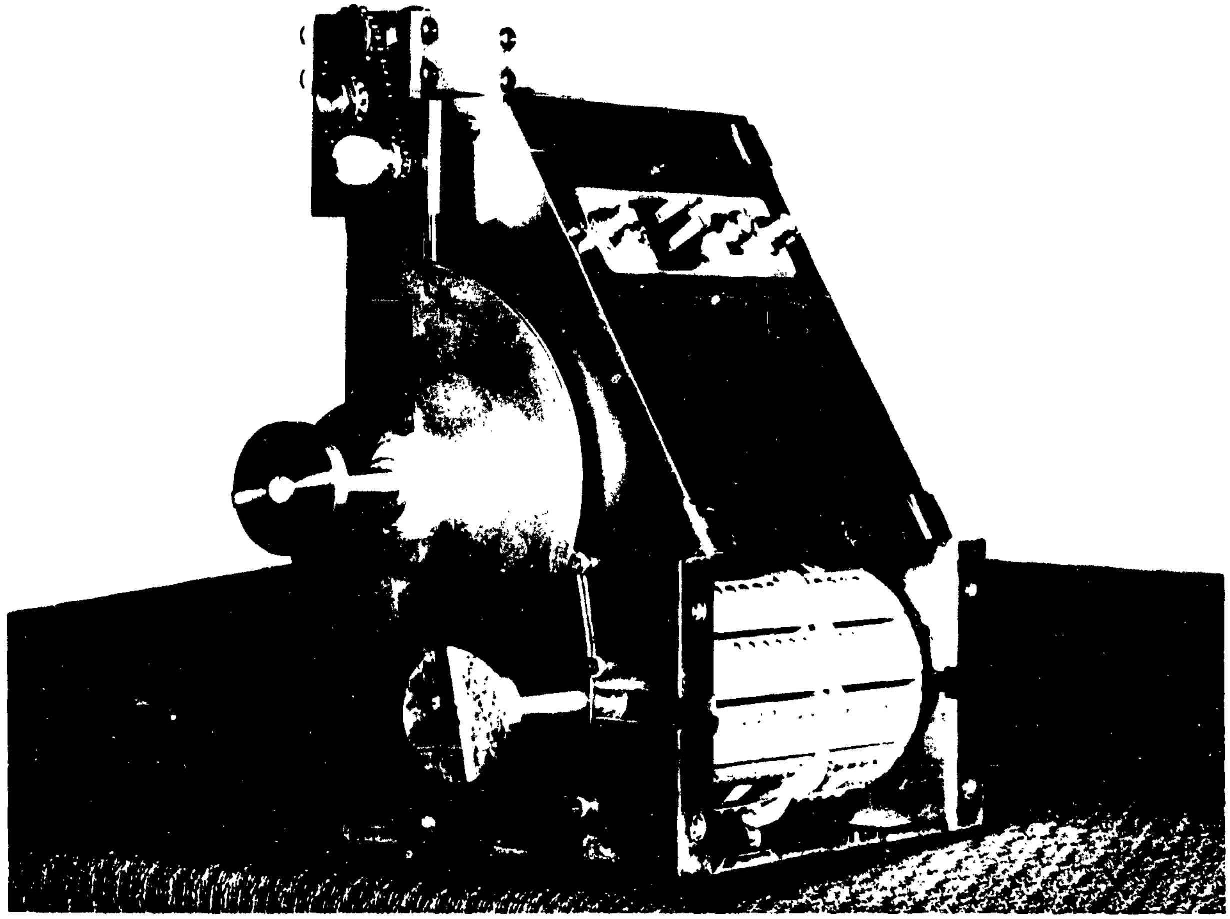


Figure 14. Tunable Phase Detector Assembly, Oblique View

ceramic capacitors then was proposed. While these were found to be excellent mechanically, they were electrically noisy, particularly with respect to low-frequency noise components. Then it was learned that a glass plunger-type variable capacitor was being made by JFD Electronics which, because of its precision fit, reduced considerably the problem of microphonics. Since its dielectric is glass, no problems were encountered with electrical noise. These glass capacitors are approximately linear with respect to plunger displacement, therefore it was necessary to use a cam-type slug rack in order to obtain a dial with a reasonably linear frequency scale. Unfortunately, these capacitors are not sufficiently similar and it was necessary to use a differential trimmer adjustment on one of the

capacitors to correct for tracking errors. This trimmer control is concentric with the main phase detector tuning control and is indicated in figure 13. The tracking problem also was aggravated by slight differences in the phase detector transformers.

The early development work on the tunable phase detector was conducted in the 1-mc to 3-mc frequency range since the most stringent resolution requirements occurred at 1 mc. Initially, the phase detector transformers were tuned with a fixed mica capacitor and an 8- to 50-picofarad ceramic trimmer capacitor. The residual peak-to-peak noise level in this circuitry remained in the range of 10 to 20 millidegrees regardless of what circuit changes were made. This condition existed until the tuning capacitors were replaced with a pair of 1- to 90-picofarad tubular glass capacitors. The residual noise level immediately dropped to approximately 1 millidegree. Substitution of various values and types of capacitors revealed that the noise-producing unit was the ceramic trimmer. No further trouble was encountered until during the final assembly work. Some silicon grease was used in the 5- to 180-picofarad capacitors in an attempt to make the phase detector tuning mechanism operate more smoothly. The grease performed the required lubricating function, but the residual noise level returned to between 10 and 20 millidegrees. Numerous washings of the capacitors were necessary using several different solvents before the noise level returned to the lower value. Subsequent investigation has shown that silicone grease was used in the assembly of the ceramic trimmers that had been used in the earlier development work.

Selection of the phase detector diodes also was found to be important. The 1N629's suggested in the study phase report were found to have too much shunt capacity for use at 110 mc. The diode capacity provides a path for r-f coupling between the signal and reference coils and thus unbalances the phase detector. As a result, low-capacity point-contact diodes were tried. Although performance was satisfactory in the vhf region, these diodes were found to be too noisy for the resolution requirements at 1 mc. The Fairchild FD1 09 planar diode then was considered. Although this diode has more capacity than the point-contact units tested, it performs satisfactorily over the entire frequency range. Two units are used in series in order to increase the maximum inverse voltage rating as well as

to reduce the shunt capacity by a factor of two. The r-f filter capacitors, C55 and C56, are chosen to present an effective r-f short and yet pass the audio-frequency output of the phase detector. With the 220-pf capacitors used, the time constant of the phase detector is approximately 2000 cps.

4.2.1.3 Mathematical Analysis of Phase Detector. As previously discussed, the 1- to 110-mc tunable phase detector is composed of three phase detectors operating individually over only a portion of the entire frequency range. Each of the circuits is merely a variation of the basic circuit shown in figure 8 and in figure 15.

In order to simplify the analysis, it is assumed that the internal impedances of the generators, Z_1 and Z_2, are small compared with R. It also is assumed that the rectifier efficiency η is a constant with respect to rectified voltage. The circuit of figure 15 then can be simplified to that of figure 16.

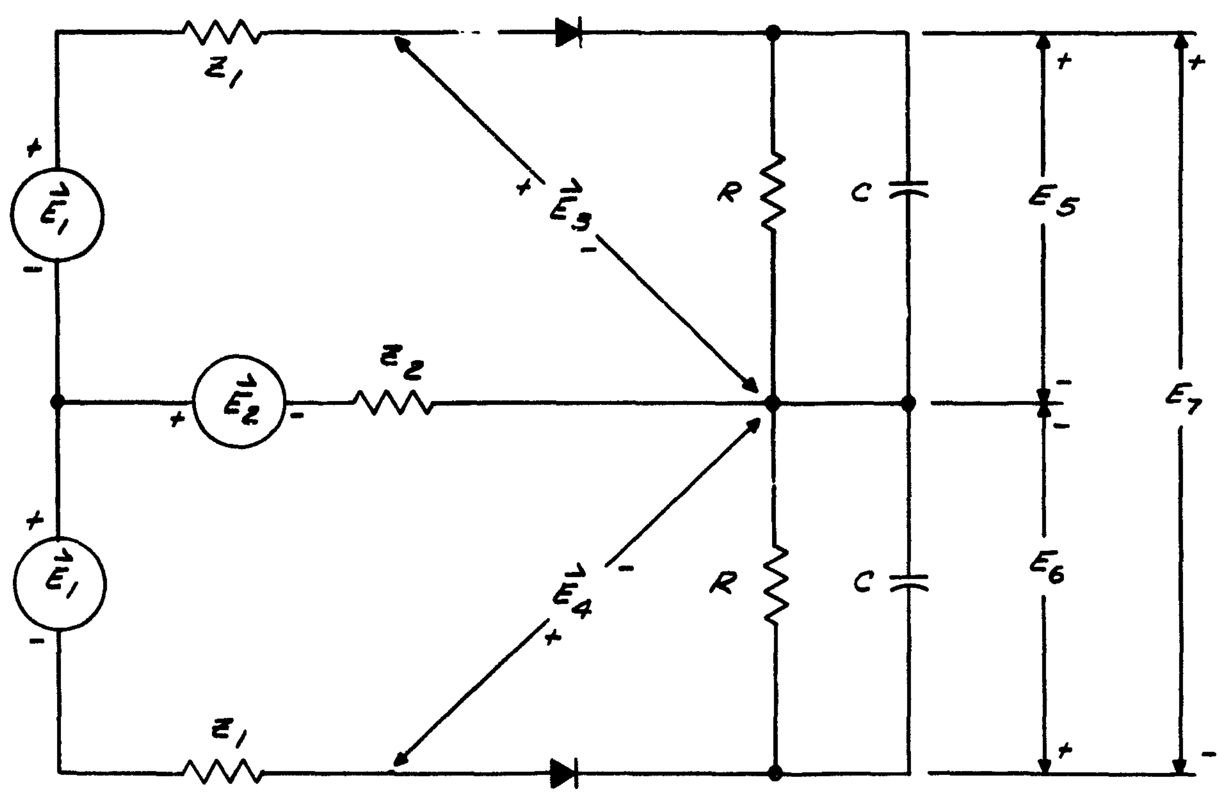


Figure 15. Phase Detector Equivalent Circuit

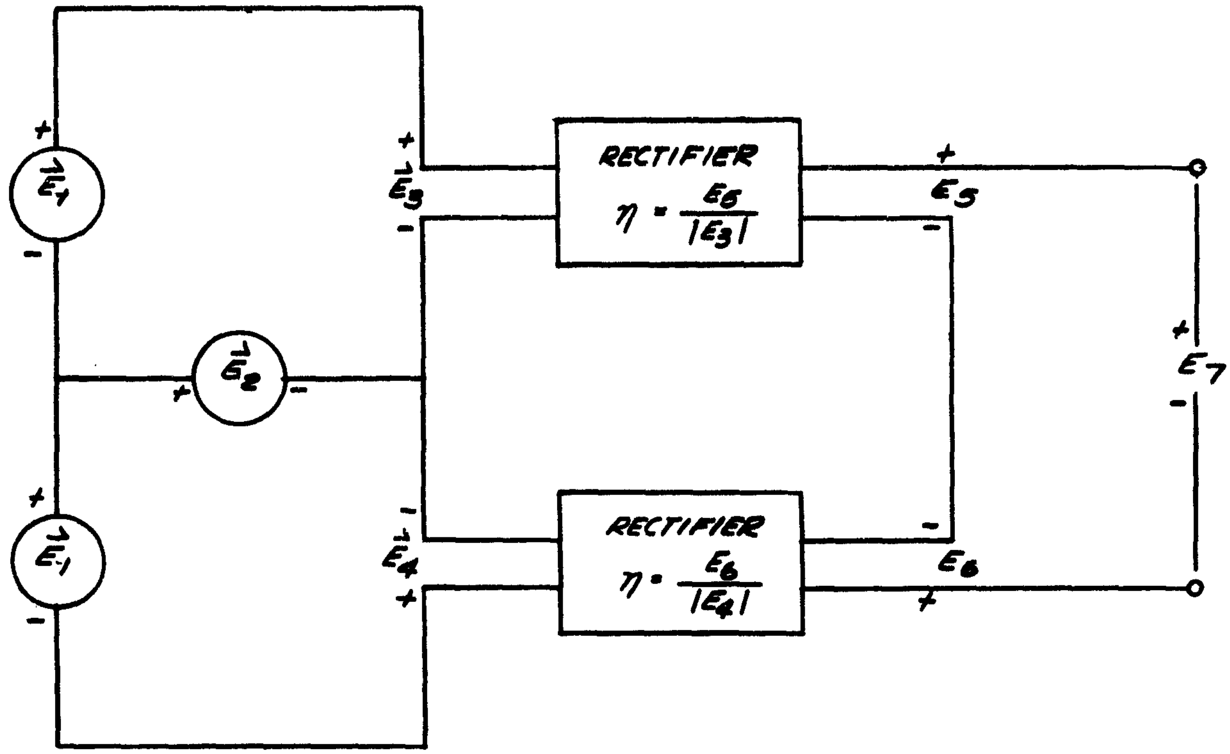


Figure 16. Phase Detector, Simplified Block Diagram

From figure 16, it can be seen that:

$$\vec{E}_3 = \vec{E}_1 + \vec{E}_2 \quad (1)$$

$$\vec{E}_4 = \vec{E}_2 - \vec{E}_1 \quad (2)$$

if
$$\vec{E}_1 = E_1 \operatorname{Sin}(\omega t + \theta) \quad (3)$$

and
$$\vec{E}_2 = E_2 \operatorname{Sin}(\omega t + \phi) \quad (4)$$

then
$$\vec{E}_3 = E_1 \operatorname{Sin}(\omega t + \theta) + E_2 \operatorname{Sin}(\omega t + \phi) \quad (5)$$

and similarly
$$\vec{E}_4 = E_2 \operatorname{Sin}(\omega t + \phi) - E_1 \operatorname{Sin}(\omega t + \theta) \quad (6)$$

These equations are shown graphically in figure 17.

From the trigonometric identities:

$$\operatorname{Sin}(\omega t + \theta) = \operatorname{Sin}(\omega t) \operatorname{Cos}\theta + \operatorname{Cos}(\omega t) \operatorname{Sin}\theta \quad (7)$$

$$\operatorname{Sin}(\omega t + \phi) = \operatorname{Sin}(\omega t) \operatorname{Cos}\phi + \operatorname{Cos}(\omega t) \operatorname{Sin}\phi \quad (8)$$

Substituting,

$$\vec{E}_3 = E_1 \left[\operatorname{Sin}(\omega t) \operatorname{Cos}\theta + \operatorname{Cos}(\omega t) \operatorname{Sin}\theta\right] + E_2 \left[\operatorname{Sin}(\omega t) \operatorname{Cos}\phi + \operatorname{Cos}(\omega t) \operatorname{Sin}\phi\right] \quad (9)$$

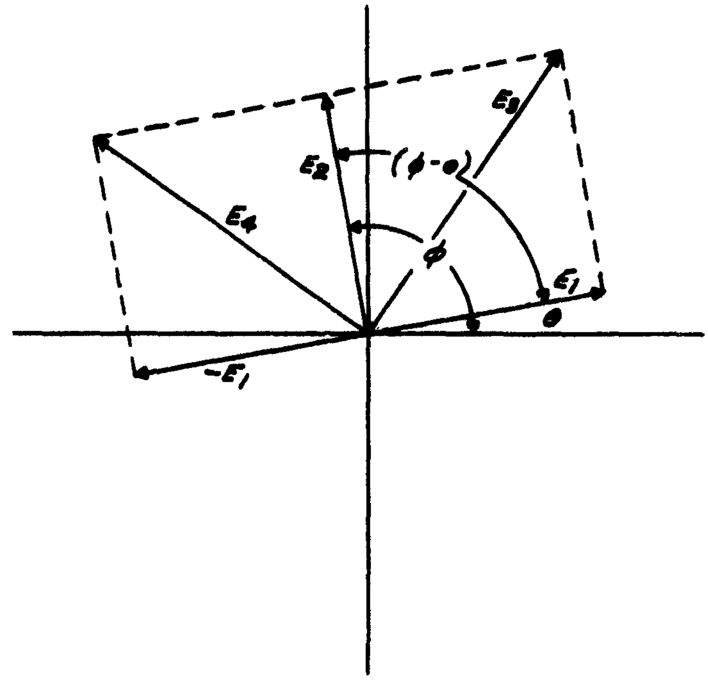


Figure 17. Phase Detector, Phasor Diagram

If E_1 is made equal to E_2 the expression can be simplified further to:

$$\vec{E}_3 = E_1 \left[\sin(\omega t)\cos\theta + \cos(\omega t)\sin\theta + \sin(\omega t) + \cos\phi + \cos(\omega t) + \sin\phi \right] \tag{10}$$

$$\vec{E}_3 = E_1 \left\{ \left[\sin(\omega t) \right] \left[\cos\theta + \cos\phi \right] + \left[\cos(\omega t) \right] \left[\sin\theta + \sin\phi \right] \right\} \tag{11}$$

$$|\vec{E}_3| = E_1 \sqrt{\left[\cos\theta + \cos\phi \right]^2 + \left[\sin\theta + \sin\phi \right]^2} \tag{12}$$

Then using the identities:

$$\cos\theta + \cos\phi = 2\cos\frac{1}{2}(\theta + \phi)\cos\frac{1}{2}(\theta - \phi) \tag{13}$$

$$\sin\theta + \sin\phi = 2\sin\frac{1}{2}(\theta + \phi)\cos\frac{1}{2}(\theta - \phi) \tag{14}$$

$$|\vec{E}_3| = E_1 \sqrt{\left[2\cos\frac{1}{2}(\theta + \phi)\cos\frac{1}{2}(\theta - \phi) \right]^2 + \left[2\sin\frac{1}{2}(\theta + \phi)\cos\frac{1}{2}(\theta - \phi) \right]^2} \tag{15}$$

$$= E_1 \sqrt{4\left[\cos^2\frac{1}{2}(\theta + \phi) \right]\left[\cos^2\frac{1}{2}(\theta - \phi) \right] + 4\left[\sin^2\frac{1}{2}(\theta + \phi) \right]\left[\cos^2\frac{1}{2}(\theta - \phi) \right]} \tag{16}$$

The expression can be simplified using the identities:

$$\sin^2\frac{1}{2}(\theta + \phi) = \frac{1 - \cos(\theta + \phi)}{2} \tag{17}$$

$$\cos^2 \frac{1}{2}(\theta + \phi) = \frac{1 + \cos(\theta + \phi)}{2} \tag{18}$$

Then

$$|\vec{E}_3| = E_1 \sqrt{[1 + \cos(\theta + \phi)]\,[1 + \cos(\theta - \phi)] + [1 - \cos(\theta + \phi)]\,[1 + \cos(\theta - \phi)]} \tag{19}$$

$$= E_1 \sqrt{1 + \cos(\theta + \phi) + \cos(\theta - \phi) + \cos(\theta + \phi)\cos(\theta - \phi) + 1 - \cos(\theta + \phi) + \cos(\theta - \phi) - \cos(\theta - \phi)\cos(\theta + \phi)} \tag{20}$$

$$= E_1 \sqrt{2 + 2\cos(\theta - \phi)} \tag{21}$$

Then using the identity:

$$\cos \frac{1}{2}(\theta - \phi) = \pm \sqrt{\frac{1 + \cos(\theta - \phi)}{2}} \tag{22}$$

$$|\vec{E}_3| = 2E_1 \left|\cos \frac{1}{2}(\theta - \phi)\right| \tag{23}$$

In a similar manner it can be shown that

$$|\vec{E}_4| = 2E_1 \left|\sin \frac{1}{2}(\theta - \phi)\right| \tag{24}$$

From figure 16 it can be seen that

$$E_5 = \eta |\vec{E}_3| = 2\eta E_1 \left|\cos \frac{1}{2}(\theta - \phi)\right| \tag{25}$$

$$E_6 = \eta |\vec{E}_4| = 2\eta E_1 \left|\sin \frac{1}{2}(\theta - \phi)\right| \tag{26}$$

$$E_7 = (E_5 - E_6) = 2\eta E_1 \left|\cos \frac{1}{2}(\theta - \phi)\right| - 2\eta E_1 \left|\sin \frac{1}{2}(\theta - \phi)\right| \tag{27}$$

$$E_7 = 2\eta E_1 \left[\left|\cos \frac{1}{2}(\theta - \phi)\right| - \left|\sin \frac{1}{2}(\theta - \phi)\right|\right] \tag{28}$$

This equation is plotted in figure 18.

The slope of the phase detector output now can be determined by differentiation of equation (28).

$$\frac{dE_7}{d(\theta - \phi)} = 2\eta E_1 \left[-\frac{1}{2}\sin \frac{1}{2}(\theta - \phi) - \frac{1}{2}\cos \frac{1}{2}(\theta - \phi)\right] \text{ for } 0 \leq (\theta - \phi) \leq 180 \tag{29}$$

$$\frac{dE_7}{d(\theta - \phi)} = -\eta E_1 \left[\sin \frac{1}{2}(\theta - \phi) + \cos \frac{1}{2}(\theta - \phi)\right] \tag{30}$$

Therefore the slope at $(\theta - \phi) = 90°$ is found to be

$$-\eta E_1 [\sin 45° + \cos 45°] = -\eta E_1 \sqrt{2} = -1.414\eta E_1 \tag{31}$$

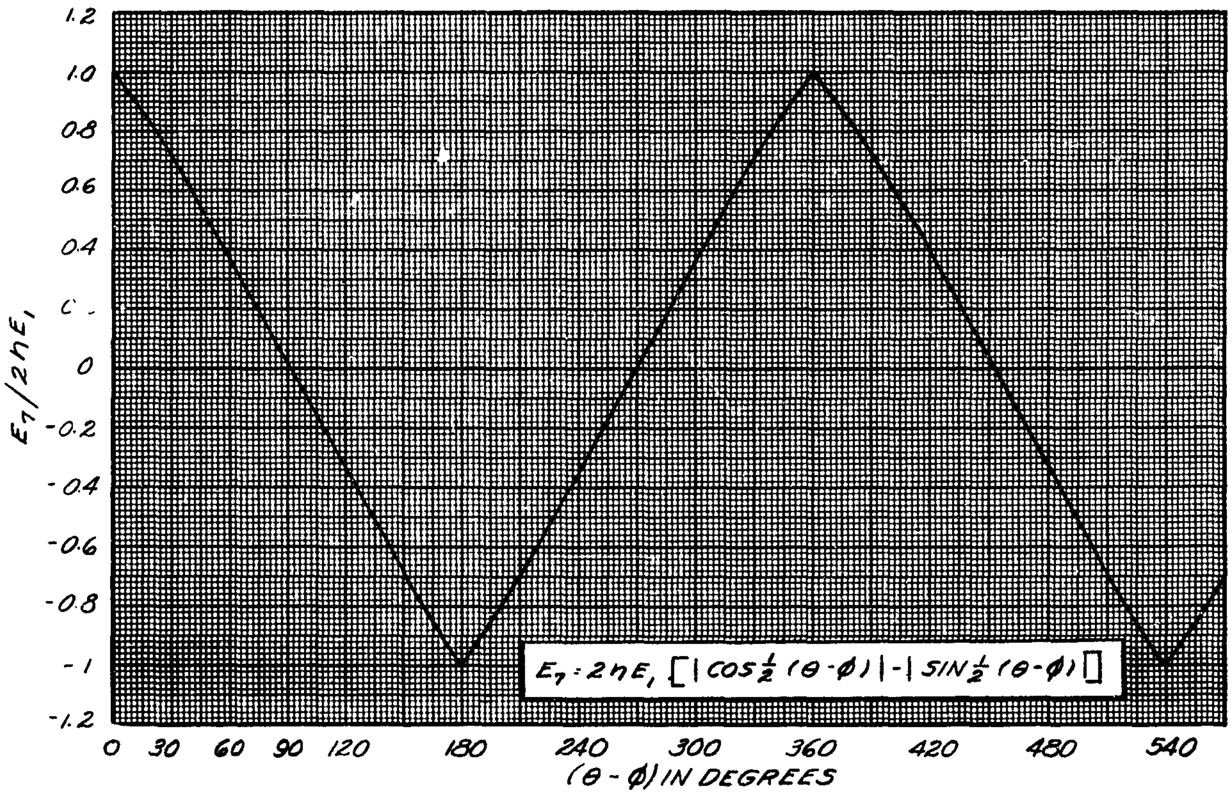


Figure 18. Phase Detector Output Versus Phase Angle

The average slope for $0 \leq (\theta - \phi) \leq 180°$ is

$$\frac{-4\eta E_1}{\pi} = -1.273\eta E_1 \tag{32}$$

The error in using the average slope at the 90-degree point is therefore

$$\frac{1.414 - 1.273}{1.414} \times 100 = \frac{14.1}{1.414} = 9.97\,\% \tag{33}$$

If the unit is calibrated using the peak-to-peak unlocked voltage, the actual value of the slope at the 90-degree point is $\frac{1.414}{1.273} = 1.11$ times greater.

4.2.2 Reference Oscillator.

4.2.2.1 General Discussion. The reference oscillator, as shown in the block diagram of figure 1, provides the second input to the phase detector. Furthermore, it is phase locked to the average frequency of the crystal-vibration-test oscillator by means of an integrating network on the output of the phase detector. The d-c voltage from this integrating network controls the voltage-variable reactance in the reference oscillator circuitry.

The reference oscillator covers the frequency range from 1 mc to 110 mc in eight bands. The hf oscillator circuit covers the range from 1 mc to 30 mc and uses bands I through IV. Bands V throu· ı VIII are utilized in the vhf oscillator circuit and cover the frequency range from 30 mc to 110 mc. Each oscillator circuit contains its separate voltage-variable reactance for completing the phase-locked loop. A single cathode follower is switched to the desired oscillator circuit by the band switch.

4.2.2.2 HF Oscillator Circuit. As previously indicated, the hf oscillator circuit covers the 1-mc to 30-mc range in the following four bands; 1 mc to 3 mc, 3 mc to 10 mc, 10 mc to 18 mc and 18 mc to 30 mc. The basic circuit used is that of a Pierce oscillator with various modifications to provide the particular performance desired.

The schematic diagram of the hf oscillator for band I is shown in figure 19. The circuit for this band is designed to cover the frequency range from 1 mc to 3 mc. On band II the oscillator covers the frequency range of 3 mc to 10 mc with a circuit identical to that of figure 19 except for the corresponding values of L_1 and L_2. Inasmuch as the resolution requirements for the phase-stability analyzer are more stringent at 1 mc, it was desirable to design the reference oscillator for maximum phase stability at this frequency. Consequently, it should be noted in figure 19 that the output is taken from the grid rather than the plate of the oscillator. This, in effect, passes the power spectrum at the plate of the oscillator through a crystal filter and limits the incidental phase modulation of the signal to the crystal bandwidth.

Since it had been decided previously (IDR-532-2) to operate all the crystals at series resonance, an inductor must be placed in series with the crystal. In order to achieve the largest operating frequency range, it was found desirable to vary this inductance as the frequency of operation was changed. This is accomplished by using high permeability powdered iron slugs in the four hf oscillator bands. In addition, best broadband operation was achieved on bands I and II with no resonant circuit employed in either the grid or plate circuit of the oscillators when using fundamental mode crystals.

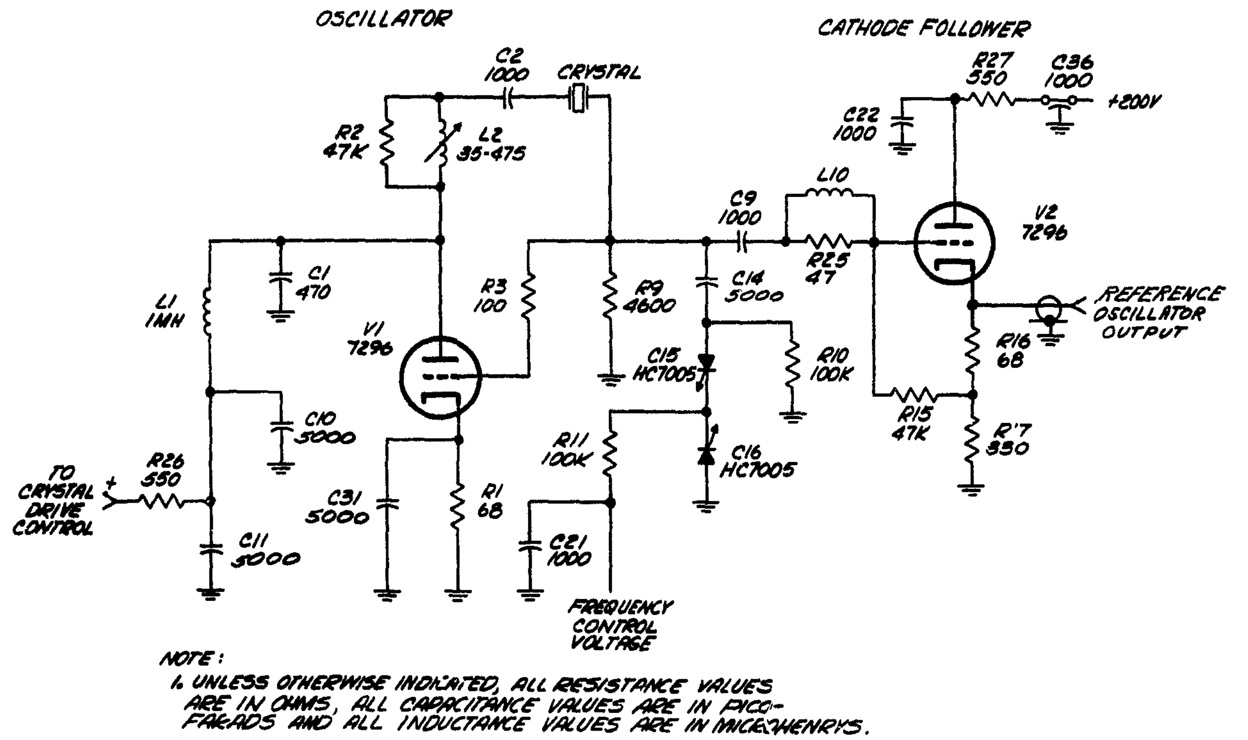


Figure 19. Reference Oscillator, Band I, 1 Mc to 3 Mc, Schematic Diagram

The desired pullability (voltage-controlled frequency excursion) for the reference oscillator on bands I and II is achieved through the use of Hughes HC-7005 Varicaps (C15 and C16). The two Varicaps are used back-to-back in order to improve the waveform. The average pullability of the oscillator on band I is approximately 2 cps/volt and 30 cps/volt on band II when using crystals with average $\frac{C_0}{C_1}$ ratios.

On bands III and IV the circuit is slightly different. A schematic diagram for band III, covering the frequency range from 10 mc to 18 mc, is given in figure 20. On band IV only the values of the equivalent parts to L_5 and L_6 are different to allow coverage of the frequency range from 18 mc to 30 mc.

For operation above 10 mc, the oscillator must be capable of operating fundamental crystals as well as third overtone units. Therefore, a provision was included to prevent third overtone crystals from oscillating at their fundamental frequency. From

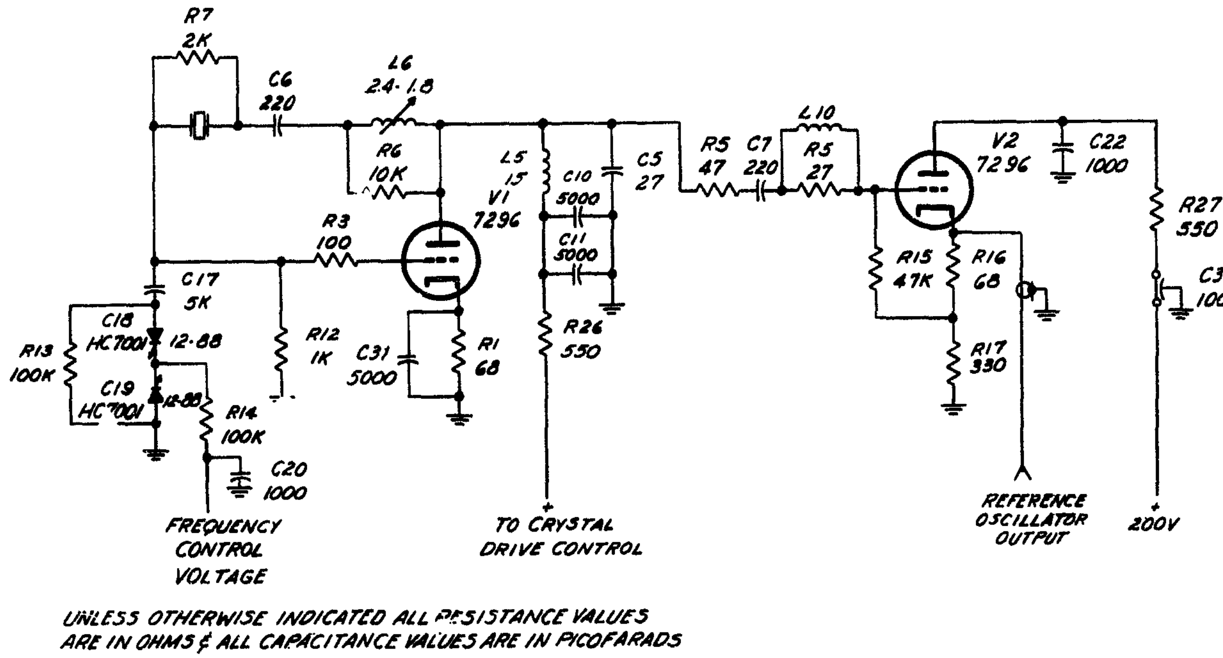


Figure 20. Reference Oscillator, Band III, 10 Mc to 18 Mc, Schematic Diagram

equation (45) of paragraph 4.2.2.5, it can be seen that g_m and R_x are positive numbers, and consequently, X_1 and X_3 must have like signs for oscillation to take place. From figure 20 it can be seen that the sign of X_1 is determined by C18 and C19 and therefore is always negative. The sign of X_3, however, is determined by the tank (L5 and C5) and can be either positive or negative. For band III, third overtone crystals have fundamental frequencies from 3.3 to 6 mc. Therefore, to ensure operation of crystals on their third overtone frequency the sign of X_3 must be positive between 3.3 and 6 mc to prevent oscillation at these frequencies. This is accomplished by making the tank resonant between 6 and 10 mc. Then, for frequencies below resonance, the tank appears inductive, the sign of X_3 is positive, and oscillation cannot take place. Components L5 and C5 are chosen as a compromise between values to give a constant capacity over the operating range and those which give a small L/C ratio to prevent unwanted oscillation through C_o of the crystal. With this tank, however, a C_o compensating network was still required. In this case, the 2000-ohm resistor (R7) connected across the crystal in figure 20 was found to be satisfactory.

In order to obtain the necessary output voltage from the reference oscillator on bands III and IV, it was found necessary to couple the cathode follower to the plate of the oscillator. This will result in somewhat higher incidental phase modulation than on bands I and II since the crystal does not act as a narrow-band filter. This is permissible, however, since the peak phase resolution requirements are lower on bands III and IV.

A different voltage-variable reactance is required on bands III and IV and Hughes HC-7001 Varicaps are used. The pullability on band III is approximately 100 cps/volt and on band IV is about 150 cps/volt.

The output voltage from the reference oscillator assembly is delivered by a cathode follower as shown in figures 19 and 20. This cathode follower is used with the hf and vhf oscillators and has a voltage gain of 0.6 and an output impedance of 64 ohms. When the crystal unit used with the reference oscillator is operated in accordance with MIL-3098B (this specification covers the maximum crystal drive level and resistance for crystal units and is summarized in appendix A), the oscillator-cathode-follower combination is capable of supplying a minimum of 0.5 volt into a 100-ohm load throughout the frequency range of 1 mc to 110 mc.

The variable inductors in series with the hf crystal required special attention. (These coils can be seen in figure 23.) Because of the close spacing between the coils, the band switch was constructed so that coils not in use are shorted. This greatly reduces the possibility of parasitic oscillations caused by stray coupling. The coil for band I uses a progressive universal-type winding and a tuning slug with a permeability of approximately 500. This winding and slug give a 13.5:1 inductance ratio. Band II also uses this high-permeability slug while bands III and IV use a slug with a permeability of 10. The coils for bands II, III, and IV are ordinary single-layer solenoids with spaced turns. The resistances shunting the variable inductors are to lower the Q of the self-resonant frequency of these coils, thereby reducing any tendency for parasitic oscillations.

In figure 4, it can be observed on the reference oscillator dial that the various frequency calibration marks are actually rectangular or a long bar. This results from the

fact that if a resistor of a value corresponding to the maximum specified crystal resistance at a particular frequency (see appendix A) is plugged into the appropriate crystal jack, the reference oscillator will tune to that frequency when the dial pointer is placed on the left-hand edge of the calibration bar. If a short is placed in the crystal socket, the reference oscillator will tune to that frequency when the dial pointer is on the right-hand edge of the calibration bar. Values of crystal resistance between the two extremes will tune to that frequency with the dial pointer within the calibration bar extremes. The frequency calibration points have been selected arbitrarily and are to serve primarily as starting points in the tuneup procedure. The final setting of the oscillator tuning will be determined by the amount a particular crystal must be pulled in order to obtain a locked-loop condition for the required measurements.

4.2.2.3 VHF Oscillator Circuit. The vhf oscillator circuit in the reference oscillator assembly covers the frequency range from 30 mc to 110 mc in four bands. The frequency range of these bands is 30 mc to 40 mc, 40 mc to 60 mc, 60 mc to 85 mc, and 85 mc to 110 mc.

The schematic diagram of the vhf oscillator on band VIII (85 mc to 110 mc) is shown in figure 21. The basic circuit is a Pierce oscillator with the main tuning element being the variable capacitor in the plate tank circuit. This circuit configuration was used because at frequencies above 30 mc, in order to prevent oscillation through C_0 of the crystal, the L/C ratio of the tank must be made so small that it makes the use of a fixed tuned tank impractical. The use of powdered iron slugs above 75 mc is precluded because of their unsatisfactory performance. The voltage-variable reactance is connected in series with the crystal.

Above 50 mc, it was found desirable to parallel the crystal with an inductor which would be antiresonant with C_0 of the crystal. This causes the motional arm equivalent circuit to appear as a simple series RCL circuit with essentially constant resistance, thereby preventing the oscillator output amplitude from changing significantly with a reactance change in the voltage-variable capacitor. Provision has been made in both the hf and vhf oscillators for plugging into the circuit a suitable inductor to be antiresonant with C_0.

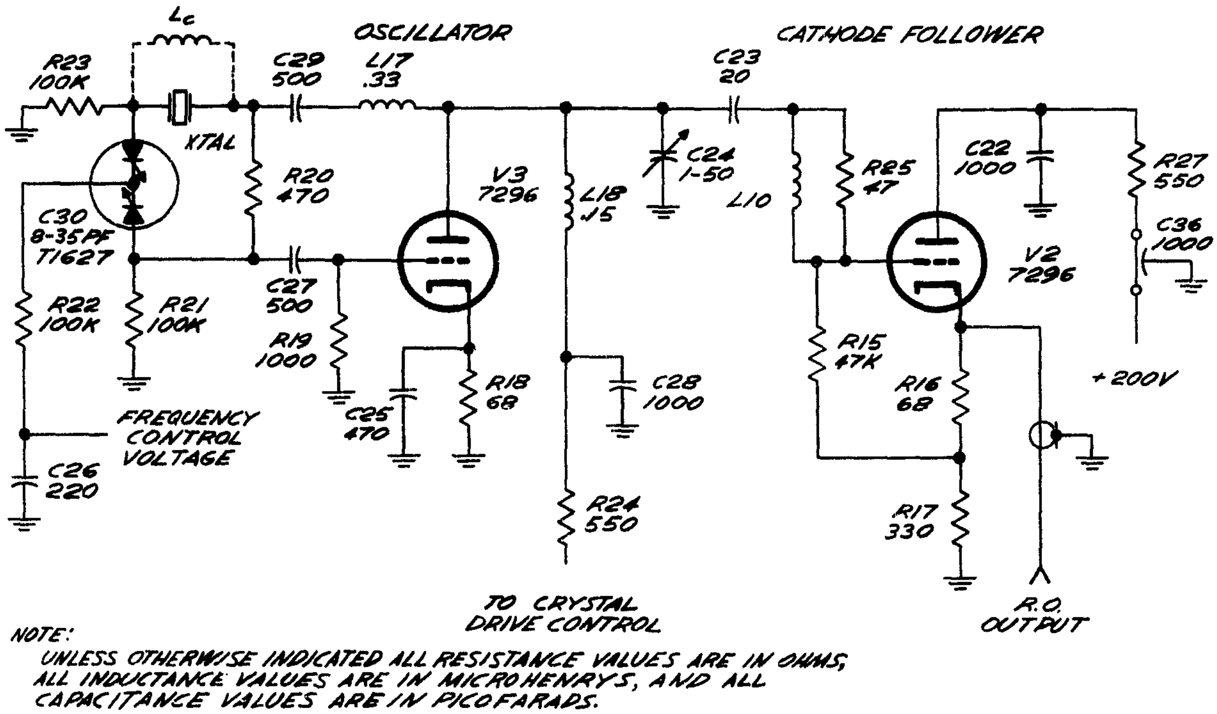


Figure 21. Reference Oscillator, Band VIII, 85 Mc to 110 Mc, Schematic Diagram

In spite of the provision for cancellation of C_0 and the use of a low L/C ratio in the tank circuit, it was found necessary to place a resistor across the crystal in order to prevent completely any oscillation through C_0. In order to prevent this resistor from reducing the crystal Q when a large voltage was placed on the voltage-variable capacitor, the resistor was connected across the series combination of the voltage-variable capacitor and the crystal.

As previously mentioned and shown in figure 21, the voltage-variable capacitor in the vhf oscillator has been placed in series with the crystal. This approach was used after difficulty was experienced in achieving proper performance with the variable reactance connected between grid and ground.

Considerable difficulty was encountered in selecting the voltage-variable reactance for use in the vhf oscillator. The voltage-variable capacitors used in the hf oscillators were found to have a very low Q at 100 mc. After suitable investigation and

testing, the Philco T-1606 Transcap was selected and used during the development work. This Transcap is a symmetrical germanium transistor-type device with a capacity range of 8 pf to 35 pf and a Q of 60 at 100 mc. However, when ordering of purchase parts for the three final models was started, it was found that Philco was no longer making the T-1606 Transcaps. An appeal to Philco for a substitute unit resulted in the use of a T-1627 Transcap. This device is quite similar to the T-1606 but is capable of providing about 3 times the pullability for the vhf oscillator.

4.2.2.4 Mechanical Design. The reference oscillator circuitry is constructed in a shielded package approximately 3 inches by 3 inches by 8 inches with a slug rack mounted on the top for tuning. Figure 22 shows an external view of the assembly. Note the crystal sockets and clips on the front surface. The pair of sockets on the left is for crystals in the 1-mc to 30-mc range and the pair on the right is for crystals in the 30-mc to 110-mc range. The lower crystal socket on each side is for HC-6 holders. Although not required by the equipment specifications, the upper crystal socket is designed for HC-25 holders and is intended to provide a socket so that an inductor of suitable value can be placed across the reference crystal in order to tune out the C_0 of that crystal. Two sets of fixed inductors encapsulated in plastic are included with each phase-stability analyzer. One set is made for use with HC-6 mounted crystals and constructed so that it plugs into the HC-25 socket. The second set is made for use with HC-18 crystals (soldered to HC-25 base) or HC-25 mounted crystals and plugs into the HC-6 socket.

The slug rack assembly is used to tune the reference oscillator. The four variable inductors used in the hf oscillator are mounted on one end of the oscillator assembly as shown in figures 23 and 24. The movement of the slugs is accomplished by a lead screw mechanism which is geared to the tuning knob on the front panel. The slugs are connected to the lifting frame by means of nylon screws. Originally, metal screws were used thereby grounding the slugs. It was discovered that small resistance changes between the metal screws and the slug rack resulted in noise modulation of the reference oscillator signal.

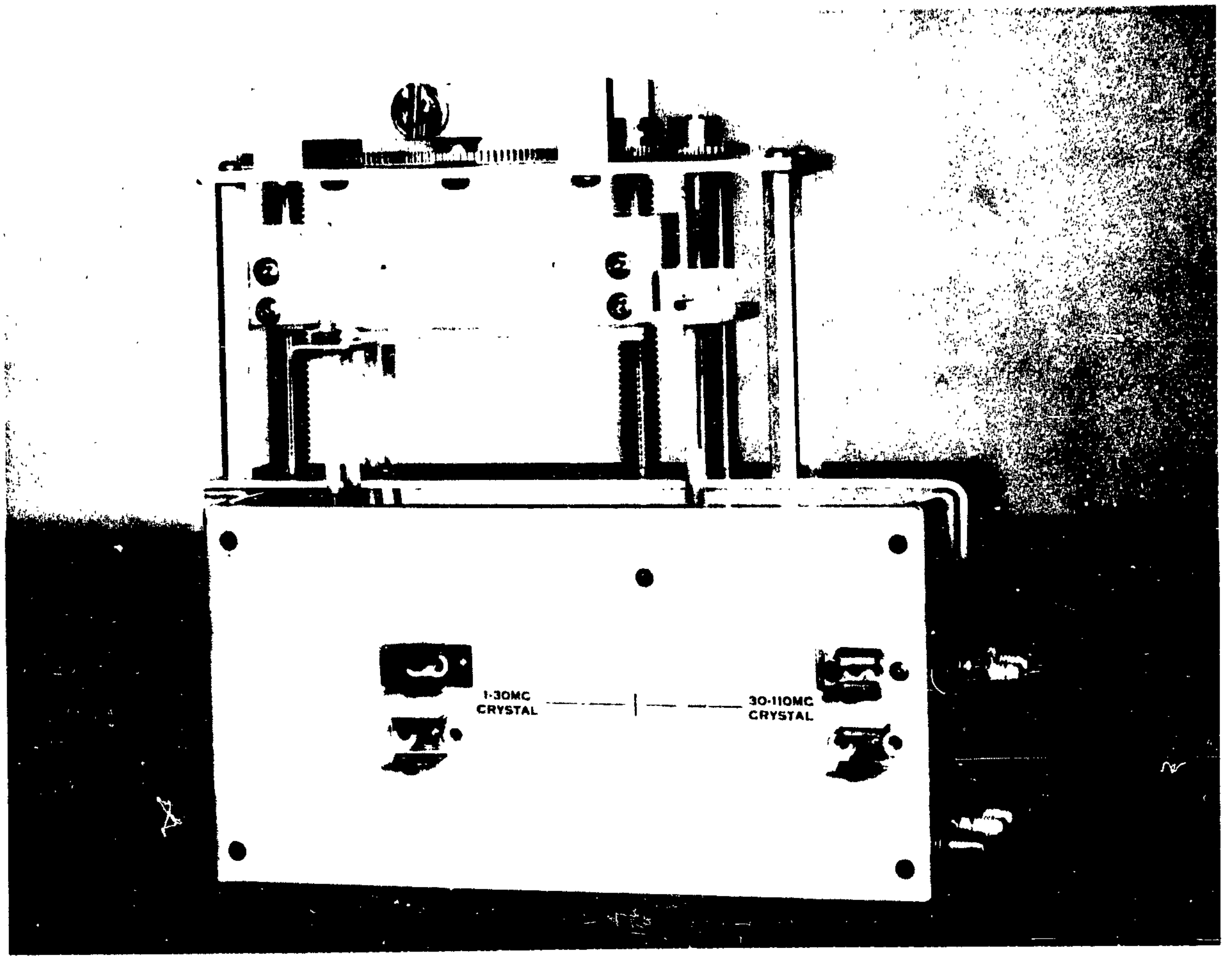


Figure 22. Reference Oscillator Assembly with Slug Rack

The vertical movement of the hf oscillator slugs is 1.5 inches while the piston travel in the vhf oscillator tubular-glass-variable capacitor is only 0.51 inch. This difference in required travel necessitated the use of a gear reduction to the third lead screw to move the capacitor piston.

As a matter of history, the initial slug rack mechanism designed for use on the reference oscillator used a lift system of bead chains. This was discarded subsequently in favor of the present lead screw mechanism because of mechanical alignment problems and the difference in travel between the coil slugs and the capacitor piston.

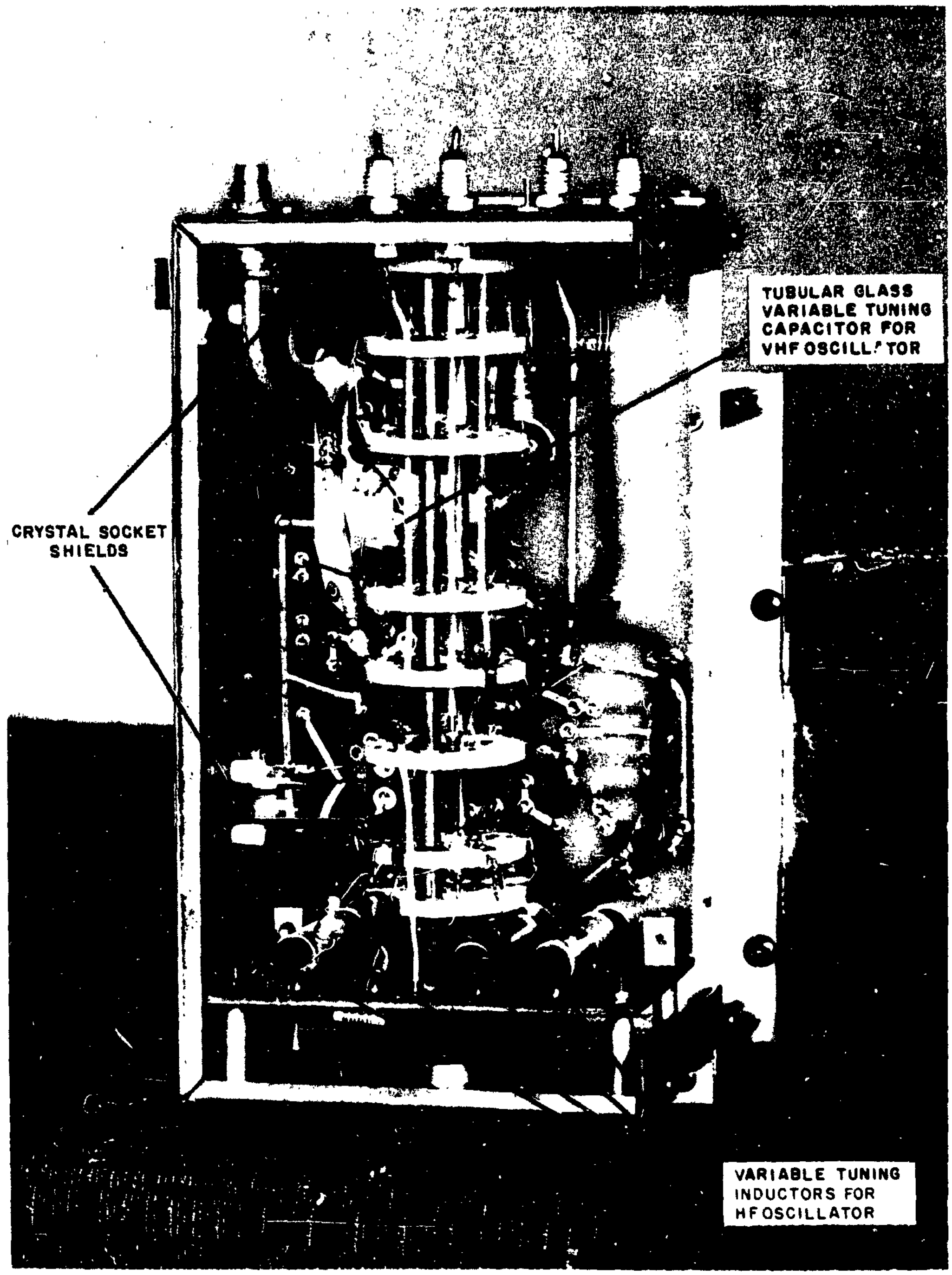


Figure 23. Reference Oscillator Assembly, Bottom View

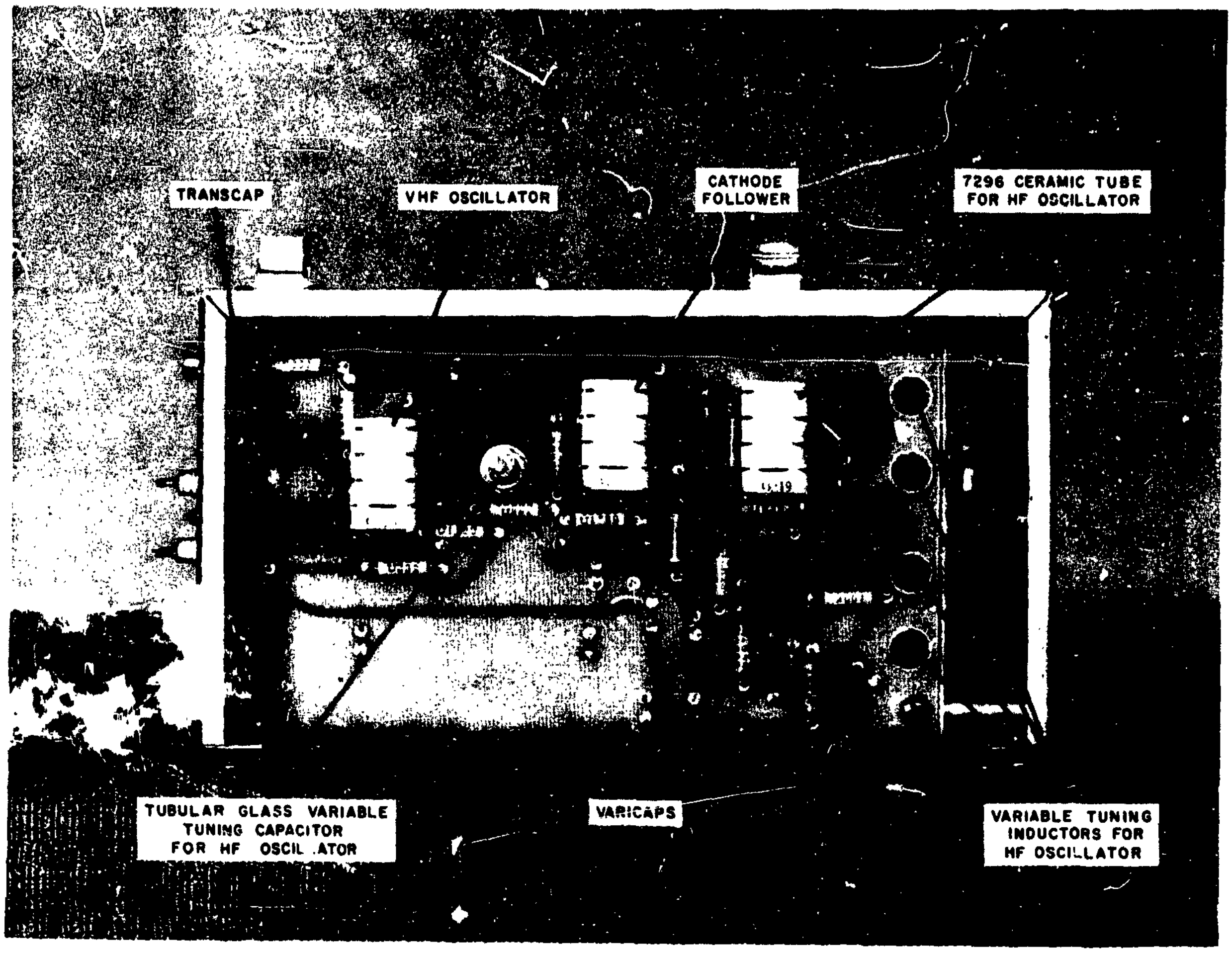


Figure 24. Reference Oscillator Assembly with Slug Rack Removed, Top View

In the layout of the printed circuit boards used in the reference oscillator, careful consideration was given to the exact placement of the art work. This was particularly important in the layout of the vhf oscillator where stray capacitance had to be kept to a minimum and r-f current ground return paths must have as low inductance as possible.

The use of the printed-circuit type ceramic tubes facilitated this design. These types of tubes were designated for use in the equipment by the design plans because of their high transconductance, resultant low-noise characteristics and good vhf performance. In addition, the rugged internal tube construction considerably reduced phase modulation effect due to microphonics. As in the case of the tunable phase detector, it

was planned originally to use the GE 7077 type ceramic tube in the reference oscillator assembly but it was found that larger dissipation ratings were necessary in order to obtain the desired output signal level. Consequently, the GE 7296 was used for the hf and vhf oscillators and the cathode follower. These tubes are shown in figure 24.

The 7296 will operate well into the vhf region, and it was found that considerable care must be taken to prevent parasitic oscillations in the region from 200 to 400 mc. These unwanted oscillations in the hf oscillator and the cathode follower were eliminated through the use of resistors and resistor-inductor parasitic suppressors in the offending sections of the circuit and by proper mechanical arrangement of the parts and printed circuitry. Specifically, a 100-ohm resistor in series with the grid of the tubes is quite effective. In the vhf oscillator, this resistor is not permissible because a series resistor combines with C_{gk} and reduces the Q of the grid circuit. In the vhf region, parasitics are prevented best by eliminating the resonant circuits which make the oscillations possible.

Other specially selected components included resistors and capacitors. Precision deposited carbon resistors are used predominantly throughout the equipment. This type of resistor has current fluctuation noise considerably below that of the standard composition carbon resistors.

Fixed capacitors in the most critical parts of the circuit are the tubular ceramic type. These appear to have lower internal noise voltages than mica and paper capacitors. Some mica types have been used in less-critical applications.

In figure 23, shield plates are shown extending from the circuit board between the pins of the two sets of crystal sockets. These shields provide a reduction in the capacitance which appears across the crystal.

The heater and high voltages are delivered to the reference oscillator assembly by individually shielded and insulated wires. The shields are grounded only at the reference oscillator thus preventing circulating ground currents from inducing unwanted voltages. The 1000-pf feedthrough-type capacitors are used to bypass these wires leading into the reference oscillator shielded compartment. The high voltage to the oscillators is variable from 10 volts to 200 volts in order to adjust the input signal to the phase detector to the required minimum of 0.5 volt.

4.2.2.5 Mathematical Analysis of a Pierce Oscillator Circuit. Preliminary work on the oscillator indicated that the best frequency stability and versatility could be obtained using the Pierce oscillator circuit. Therefore, a thorough mathematical investigation of this circuit was conducted. The basic circuit is shown in figure 25.

To simplify the analysis, the following initial assumptions are made:

1. $\frac{1}{\omega C_k} << R_k$ (The cathode resistor is bypassed.)
2. $\frac{1}{\omega C_1} << Z_3$
3. $R_g >> Z_1$
4. Z_1 is entirely reactive and includes C_{gk} of the tube.
5. Z_3 contains r_p and C_{pk} of the tube.
6. Z_2 contains C_{pg} of the tube as well as the crystal-holder capacitance.

The a-c equivalent of figure 25 is shown in figure 26, where R_x is the crystal resistance if C_{pg} is negligible. R_3 is the parallel combination of the load resistance and r_p. The circuit of figure 26 can be redrawn in the form of figure 27.

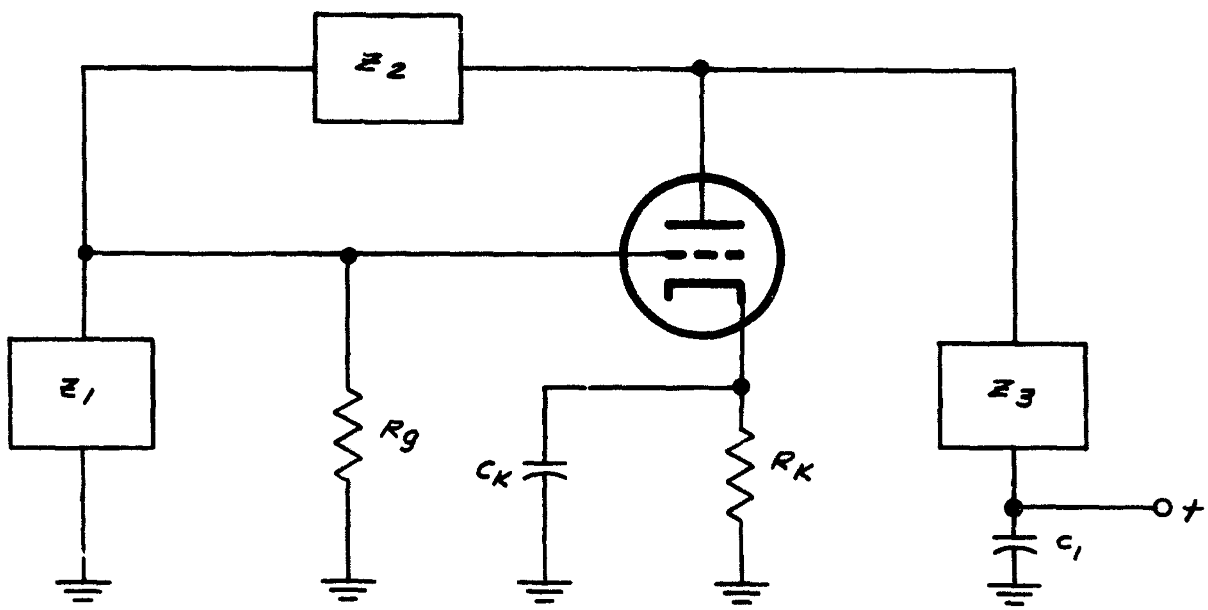


Figure 25. Basic Pierce Oscillator Circuit

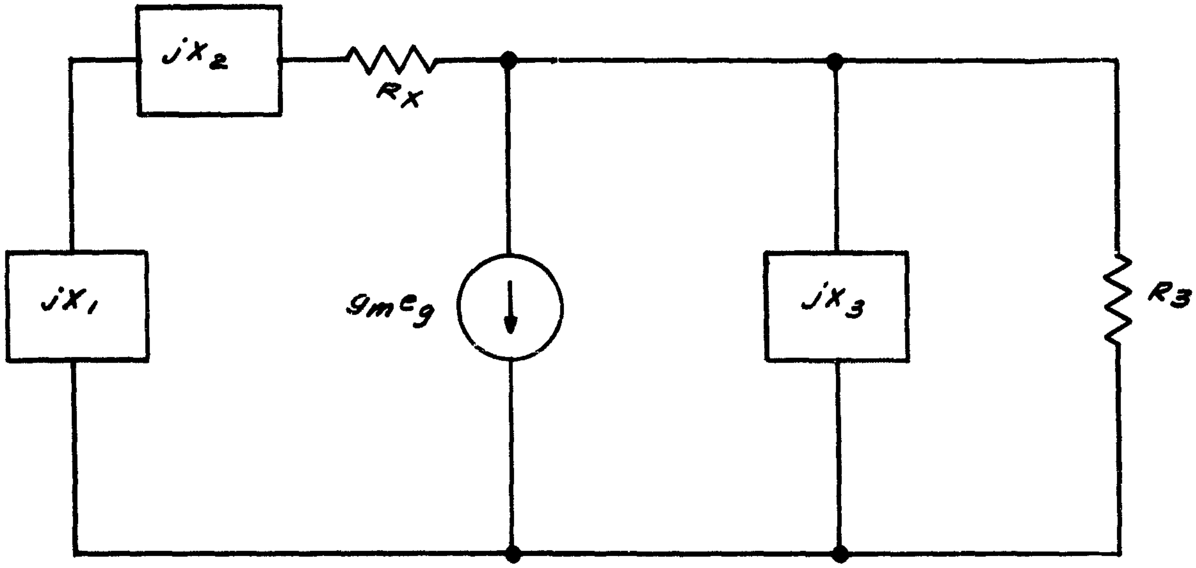


Figure 26. Equivalent Pierce Oscillator Circuit

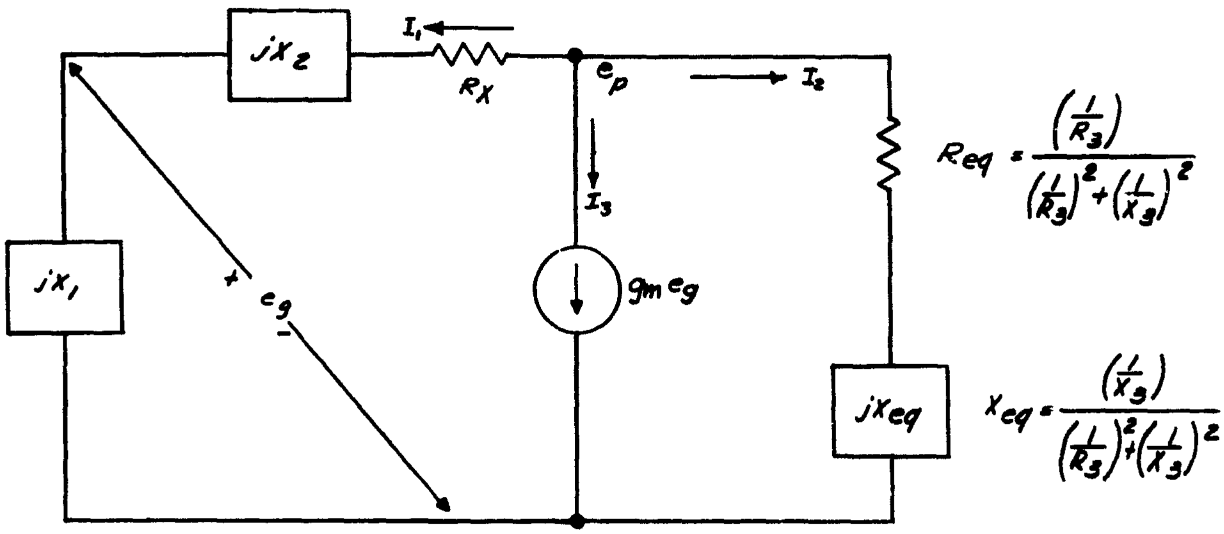


Figure 27. Simplified Equivalent Pierce Oscillator Circuit

From Kirchhoff's current law we may write:

$$I_1 + I_2 + I_3 = 0 \tag{34}$$

$$\frac{e_p}{R_x + j(X_1 + X_2)} + \frac{e_p}{R_{eq} + jX_{eq}} + g_m e_g = 0. \tag{35}$$

$$\text{But,} \quad e_g = \frac{j e_p X_1}{R_x + j(X_1 + X_2)} \tag{36}$$

Therefore,

$$\frac{e_p}{R_x + j(X_1+X_2)} + \frac{e_p}{R_{eq} + jX_{eq}} + \frac{j g_m e_p X_1}{R_x + j(X_1 + X_2)} = 0 \tag{37}$$

and

$$\frac{e_p + j e_p \; g_m X_1}{R_x + j(X_1 + X_2)} = -\frac{e_p}{R_{eq} + jX_{eq}} \tag{38}$$

Then dividing by e_p, we obtain:

$$\frac{1 + j g_m X_1}{R_x + j(X_1 + X_2)} = -\frac{1}{R_{eq} + jX_{eq}} \tag{39}$$

$$-R_x - j(X_1 + X_2) = R_{eq} + j\,X_{eq} + jR_{eq}X_1 g_m - g_m X_1 X_{eq} \tag{40}$$

This complex equation can be reduced to two equations by equating the real parts and the imaginary parts.

$$\text{Real parts:} \quad -R_x = R_{eq} - g_m X_1 X_{eq} \tag{41}$$

$$g_m X_1 X_{eq} = R_x + R_{eq} \tag{42}$$

$$\text{Imaginary parts:} \quad -(X_1 + X_2) = X_{eq} + R_{eq} X_1 g_m \tag{43}$$

$$X_1 + X_2 + X_{eq} = -g_m X_1 R_{eq} \tag{44}$$

In most applications $R_x >> R_{eq}$, $g_m R_{eq} << X_1$, and $R_3 >> X_3$. For these conditions, equations (42) and (44) can be reduced to equations (45) and (46), respectively. These are the equations which must be satisfied for oscillation to take place.

$$g_m X_1 X_3 = R_x \tag{45}$$

$$X_1 + X_2 + X_3 = 0 \tag{46}$$

4.2.3 Crystal-Vibration-Test Oscillator.

4.2.3.1 General Discussion. The crystal-vibration-test oscillator (CVTO) is used in conjunction with the quartz crystal unit under vibratory test and supplies the second input signal to the phase detector. The design plans referred to this circuit as the shake oscillator because it was originally to be mounted on the vibration table. It was planned to use 7296 ceramic vacuum tubes on separate printed circuit boards for each required frequency range. After considerable effort had been expended on designing, developing, and constructing an assembly for test on the vibration table, it was found that even the rugged construction of the ceramic tubes did not prevent phase modulation of the output signal. As a consequence, the circuitry was removed from the vibration table and the multiband oscillator assembly was renamed the crystal vibration test oscillator. For brevity, it will be referred to as the test oscillator.

Placing the test oscillator near the vibration table necessitated operating the crystal at the end of a section of double coaxial transmission line. This presented a number of problems in the design of a circuit which would function satisfactorily with the crystal unit so located. Below approximately 15 mc, no difficulty was encountered in using a phase shift analysis in the design of a circuit. A variable delay line was obtained from ECS Corporation and was tested in the oscillator circuit shown in figure 28. This circuit performed very well to about 15 mc, which was the upper frequency limit of the delay line.

Between 15 mc and 110 mc the problem of operating a crystal at the end of a double coaxial line becomes much more complex. At a particular frequency, the use of one-half wavelength of line on each side of the crystal is required. However, it is extremely difficult, if not impossible, to have the circuit operate properly over the entire frequency range with a fixed length of line without complicated circuitry at each end of the transmission line. Therefore, a different length of coaxial line was used, according to the various frequency range segments. Considerable experimental work was conducted in an effort to develop an optimum circuit. The end result of this work was to use a modified

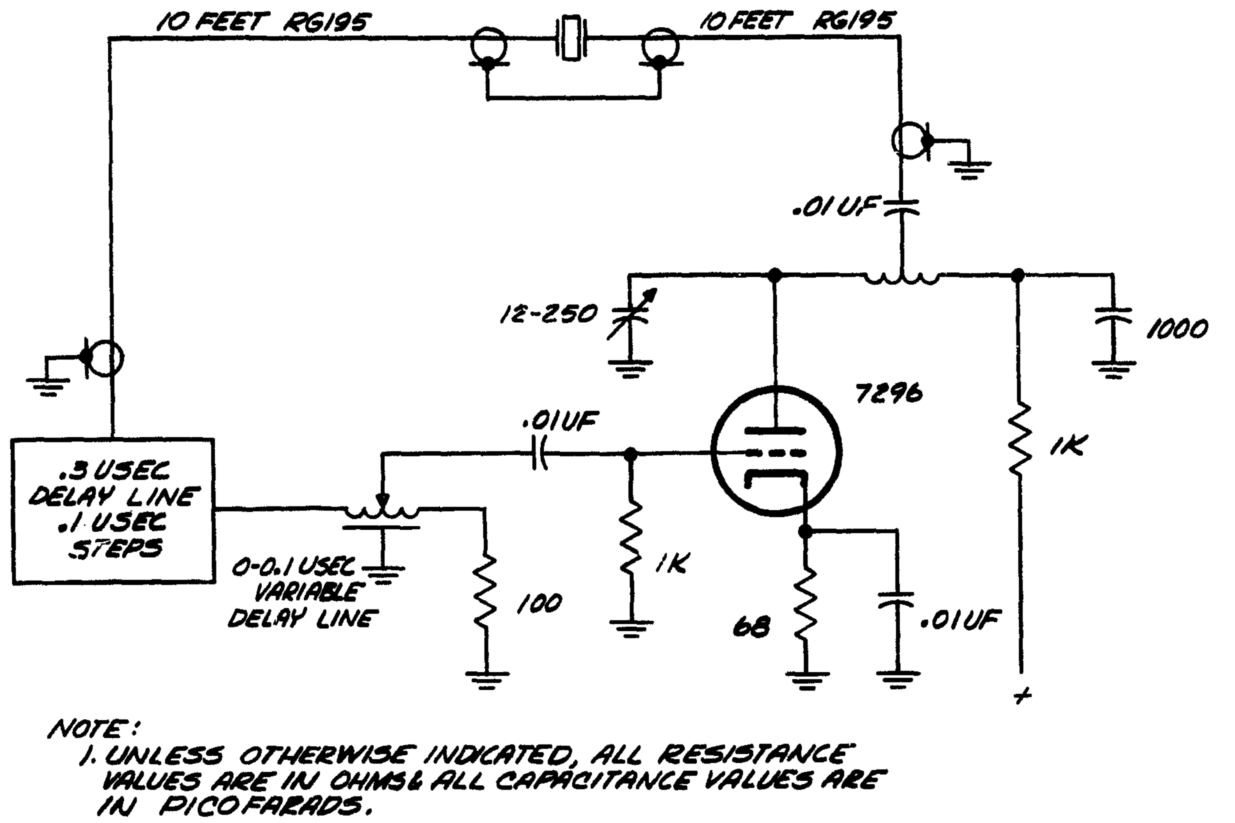


Figure 28. 1.0-Mc to 15-Mc Delay Line Oscillator, Schematic Diagram

version of the reference oscillator. The schematic diagram of figure 29 shows the entire circuit for the test oscillator. This diagram is simplified in the one-band circuits for the hf and vhf oscillators in figures 30 and 31 respectively. It should be noted that the voltage variable capacitors have been replaced with fixed capacitors. Also, the hf oscillator variable inductors L51, L53, L55, and L57 have slightly different values from those used in the reference oscillator. This was necessary in order to provide satisfactory overlap in frequency coverage for each length of coaxial transmission line to the crystal.

4.2.3.2 Mechanical Design. A photograph of the crystal-vibration-test oscillator is shown in figure 32. The assembly is housed in a cabinet 9-1/16 by 6 by 11-9/16 inches and weighs 9 pounds. The cabinet has been covered with a scuff-resistant vinyl paint. A contemporary-styled handle is provided on the top for carrying the unit, and rubber feet are fastened on the bottom.

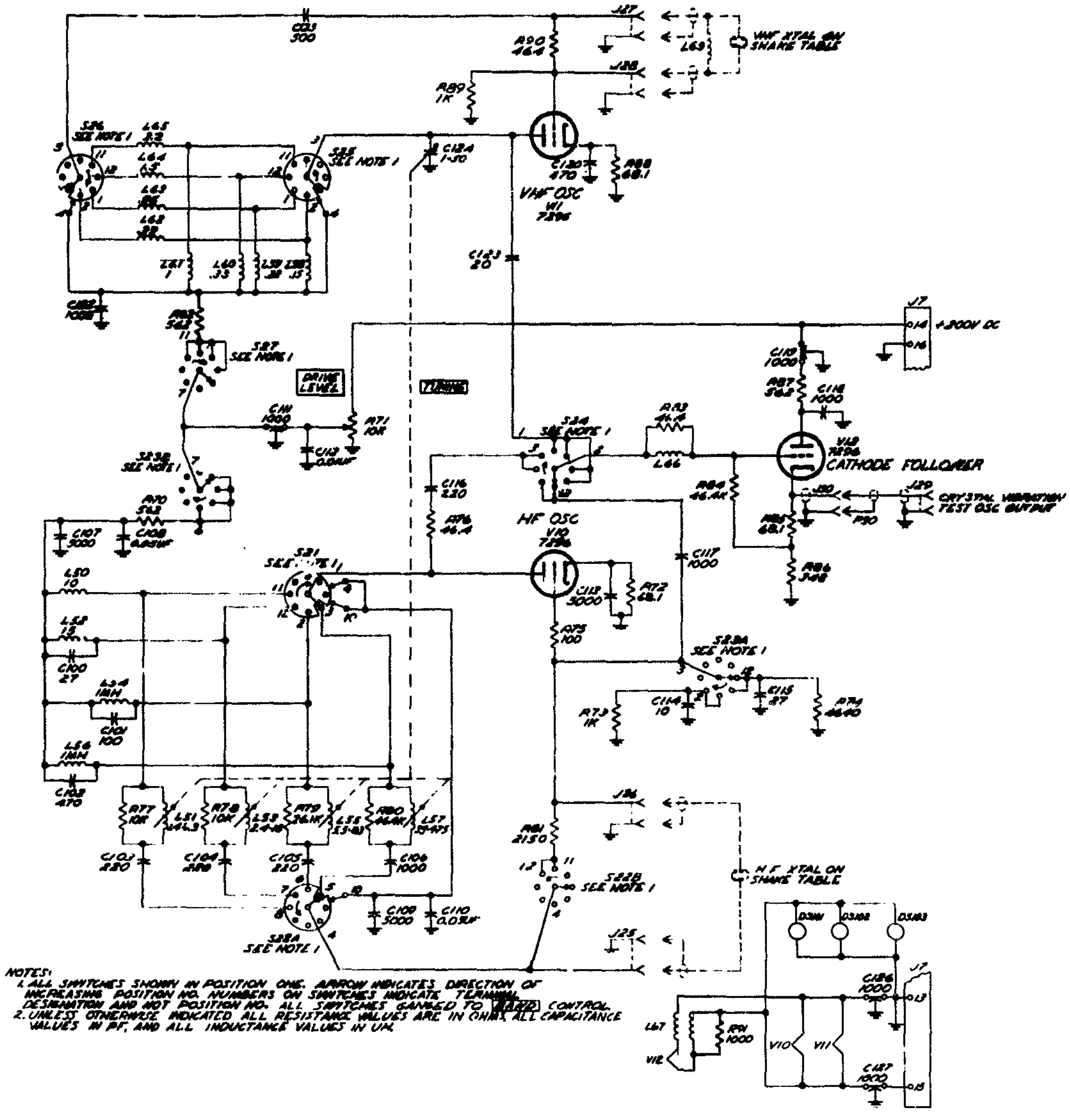
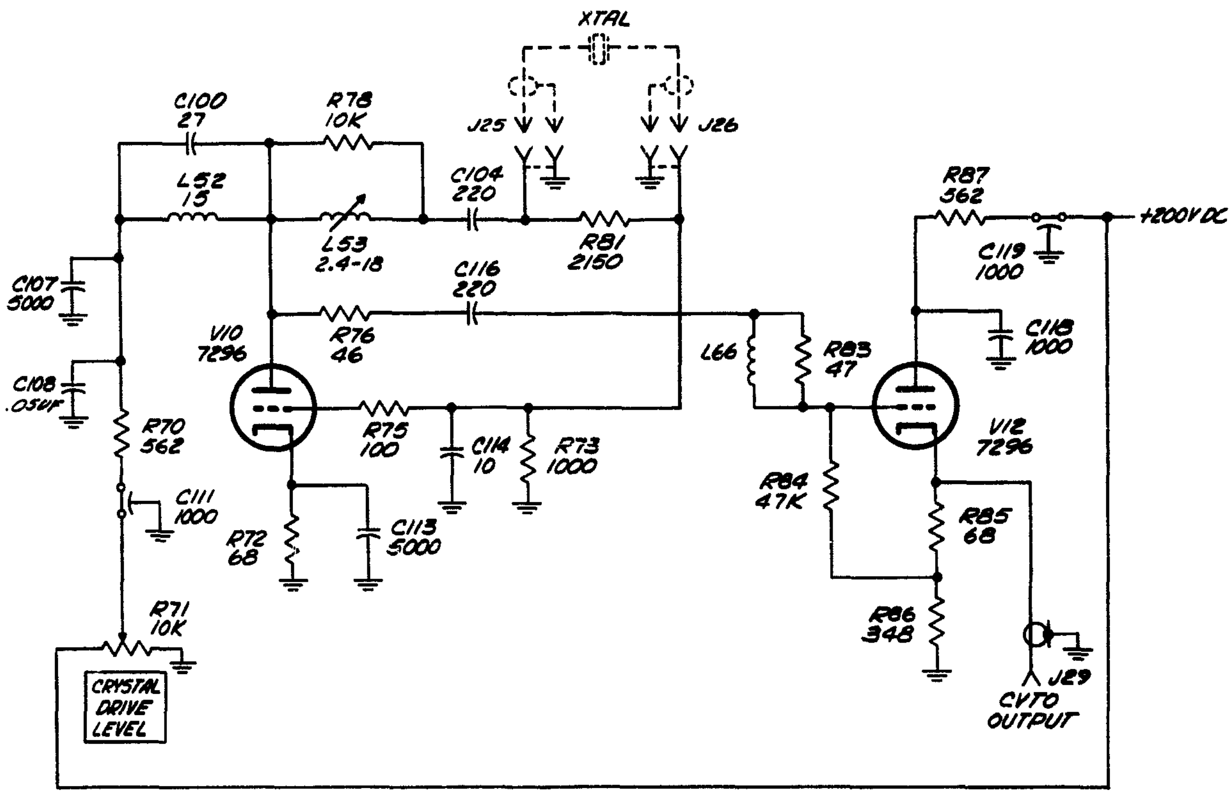


Figure 29. Crystal-Vibration-Test Oscillator, Schematic Diagram



NOTE:
1. UNLESS OTHERWISE INDICATED ALL RESISTANCE VALUES ARE IN OHMS, ALL CAPACITANCE VALUES ARE IN PICOFARADS AND ALL INDUCTANCE VALUES ARE IN MICROHENRYS.

Figure 30. Crystal-Vibration-Test Oscillator, Band III, Schematic Diagram

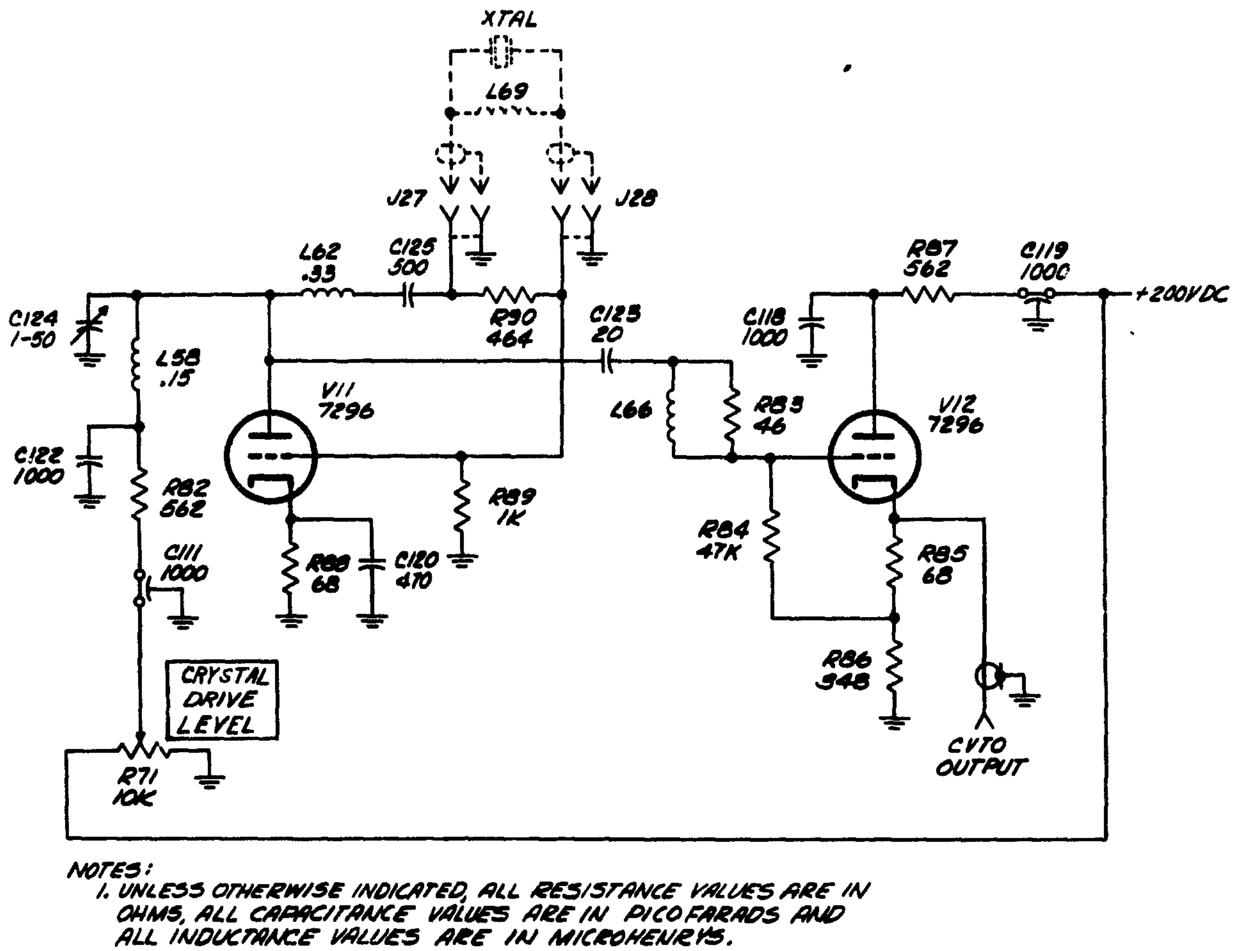


Figure 31. Crystal-Vibration-Test Oscillator, Band VIII, Schematic Diagram

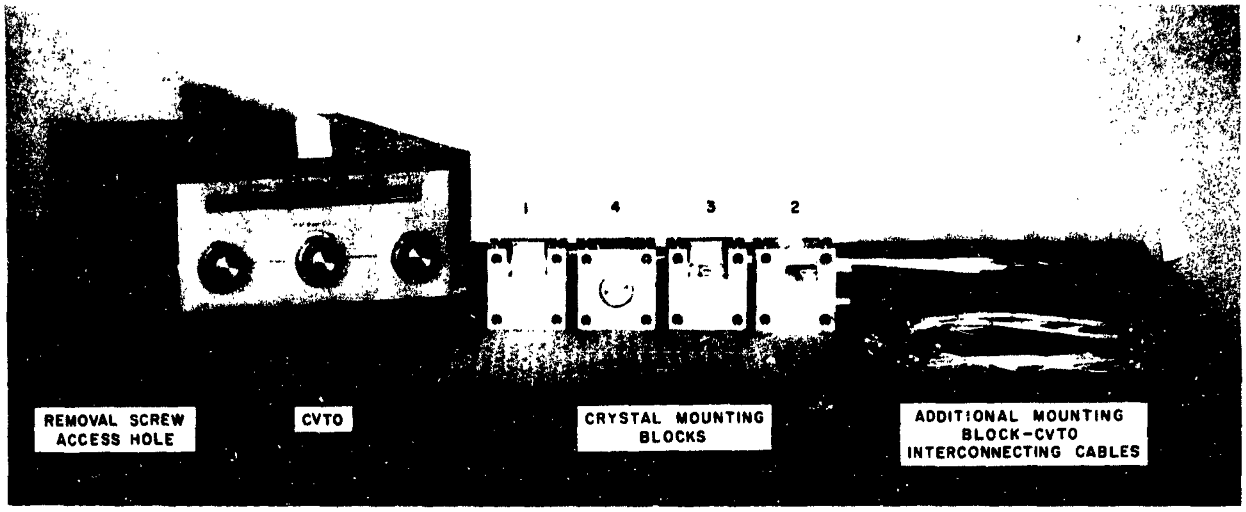


Figure 32. Crystal-Vibration-Test Oscillator and Accessories

The four crystal mounting blocks for use with the CVTO also are shown in figure 32. These blocks are designed to provide a rigid mount, free from mechanical resonances, for the various types of crystals which must be tested. The blocks are made from cold-rolled steel to allow the use, where desired, of a magnetic hold-down on the vibration table. The blocks are zinc-plated and chromate-dipped to prevent oxidation. To each mounting block are permanently attached two coaxial cables 26 inches long. These cables are a low-noise type which has a conductive powder in the shield under the outer insulation cover. This conductive powder reduces to a negligible level any noise voltages generated by the flexing of the shield wires of the coaxial cable during vibration. These cables fasten to the appropriate connectors on the top of the CVTO. The individual blocks are numbered at each end of the coaxial cables and are used with the different crystal types as follows:

Block number 1 is for use with an HC-6/U type holder containing a crystal above 50 mc, necessitating the use of a compensating inductor across the crystal leads.

Block number 2 is for use with crystals mounted in a T-5-1/2 bulb. The crystal may be mounted in the block using a potting compound such as dental investment or, as suggested by USASRDL, melted paraffin. For crystals above 50 mc a compensating inductor should be soldered across the coaxial cable terminals inside the block.

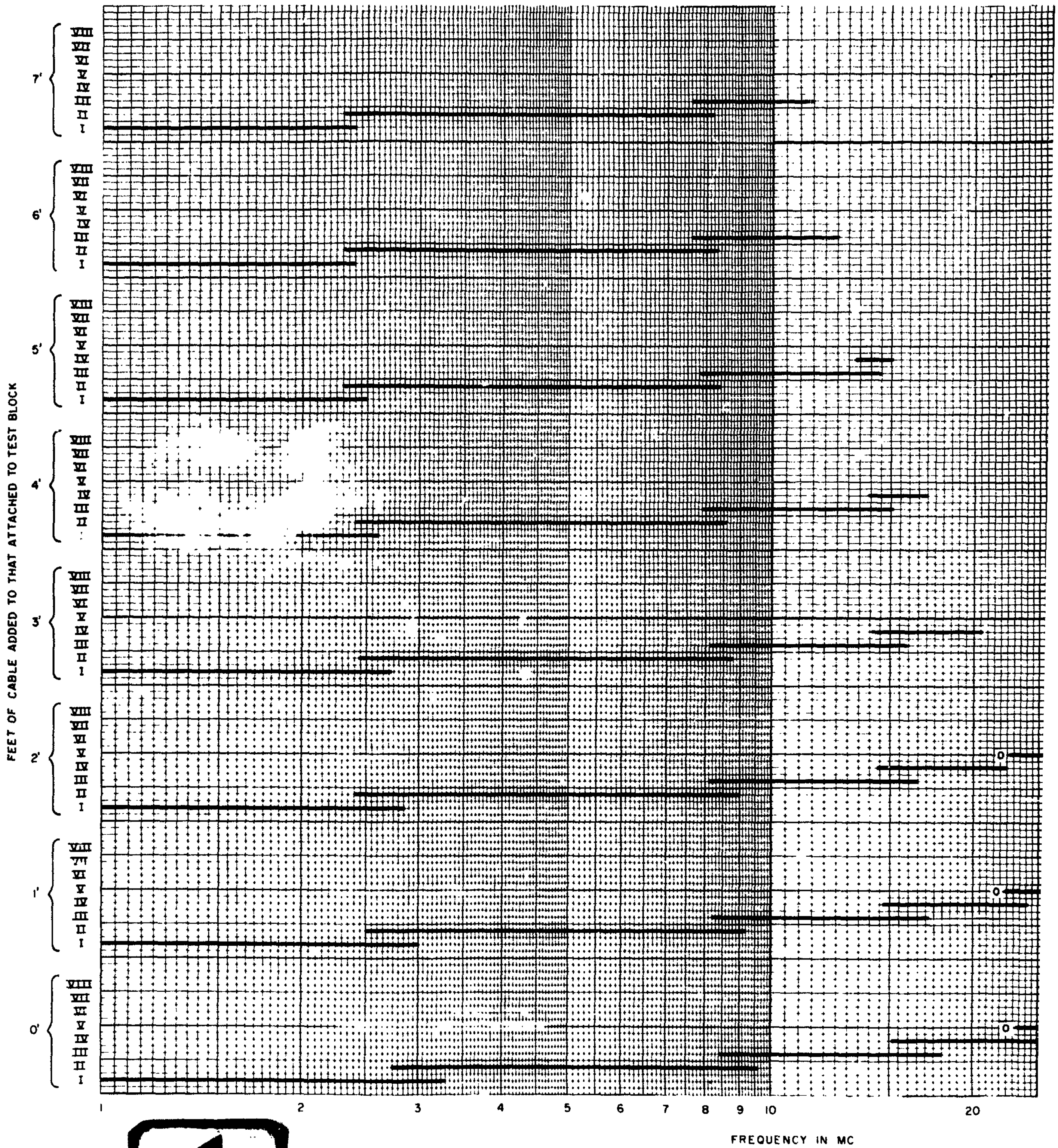
Block number 3 is used with HC-18/U crystal holders and is designed to provide for mounting of the compensating inductor across the coaxial cable terminals inside the block.

Block number 4 is used with HC-6/U holders containing crystals below 50 mc where no compensating inductor is required. The crystal plugs into the recess under the threaded cover plug.

The additional pieces of coaxial cable in figure 32 are 1-, 2-, and 4-foot lengths of RG-196/U miniature coaxial cable, which can be attached to the low-noise cable on the mounting blocks. These extra lengths of cable are necessary to provide satisfactory operation above approximately 30 mc. The selection of the proper length of cable to be used at a particular frequency is accomplished with the use of figure 33. For bands V through VIII, this graph also gives the dial numbers between which satisfactory operation of the CVTO can be expected with a particular length of cable. Because of the change in frequency range with different lengths of coaxial line, the dial of the CVTO for bands V through VIII has 0 to 100 logging scales only. The logging scale numbers are those shown in figure 33. On bands I through IV, the dial is calibrated in the same manner as the reference oscillator.

The reference oscillator chassis of figure 22 has been modified mechanically in order to mount in the test oscillator chassis as shown in figure 34. The crystal sockets have been replaced with the threaded coaxial connectors shown on the top surface of the test oscillator chassis. The switch shaft has been extended directly to the BAND knob on the front panel. A bead chain is used to turn the dial drum. The TUNING knob shaft has been rotated 90 degrees from its position in the reference oscillator assembly. The power plug receptacle and the coaxial output connector are located on the rear skirt of the chassis. The right-hand side of the chassis has been perforated with a pattern of small holes in order to match the holes in the test oscillator chassis and provide some convection ventilation past the ceramic tubes.

The test oscillator chassis is mounted in the case by means of three special neoprene isolators. The chassis must be disconnected from these mounts by removal of


FEET OF CABLE ADDED TO THAT ATTACHED TO TEST BLOCK
7'
6'
5'
4'
3'
2'
1'
0'
I
II
III
IV
V
VI
VII
VIII
1
2
3
4
5
6
7
8
9
10
20
FREQUENCY IN MC

1

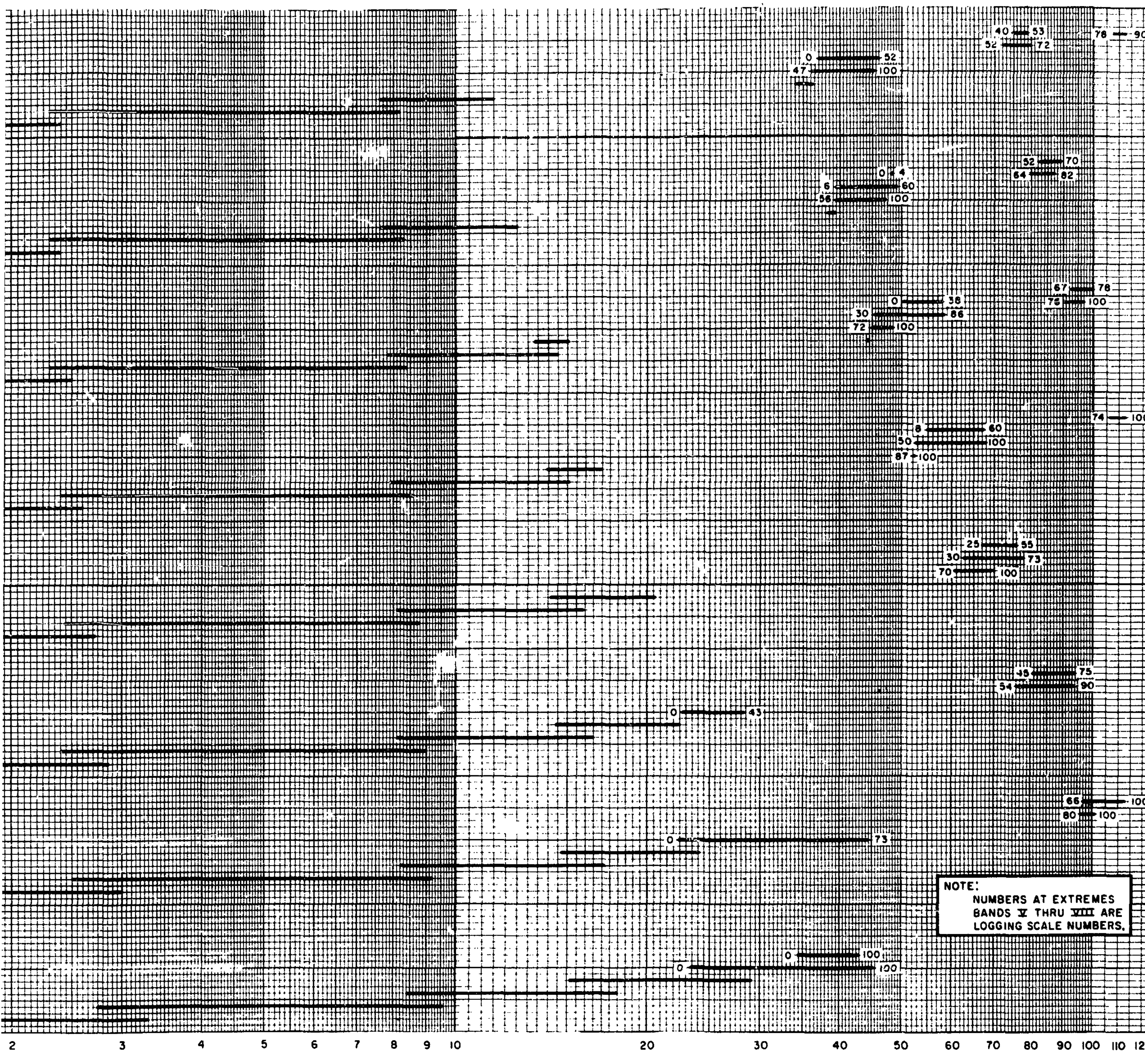


Figure 33. Test Block and CVTO Interconnecting Cable, Cable Length Vs Frequency

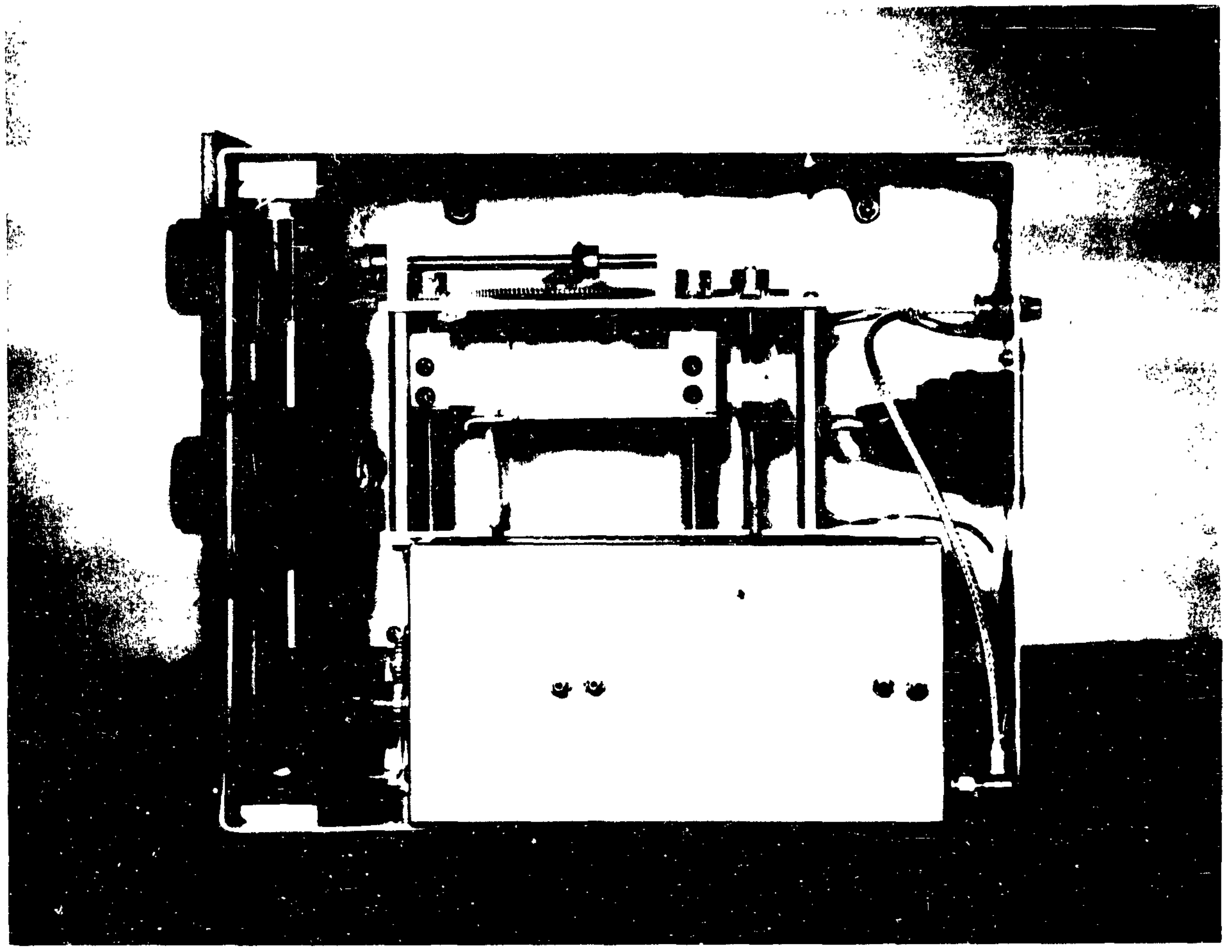


Figure 34. CVTO Removed from Case

the screws, access to which is through the holes in the case. One typical access hole is indicated in figure 32.

4.2.3.3 Analysis of Crystal-Vibration-Test Oscillator. The addition of a length of coaxial cable to connect each side of the quartz crystal to an oscillator circuit greatly increases the complexity of a complete circuit analysis. As an aid to the understanding of this circuit arrangement, the following material was developed.

Referring to paragraph 4.2.2.5 of this report on the Mathematical Analysis of a Pierce Oscillator Circuit, it is seen that two conditions must be satisfied in order that sustained oscillation will occur:

(1) $X_1 + X_2 + X_3 = 0$ (47)

This means that if X_1 and X_3 are capacitive, X_2 must be inductive and vice versa. Thus, $X_1 + X_3 = X_2$.

(2) $R_x = X_1 X_3\ g_m$ (48)

Further demonstration of these equations may be made graphically, as in figures 35, 36, and 37.

In figure 35, the sum of X_1 and X_2 as a negative quantity is plotted as a function of frequency. At any particular frequency, the ordinate represents the value of X_3 that the circuit must contain in order that oscillation at that frequency will result. The series resonant frequency of X_1 and X_2 is represented by f_o.

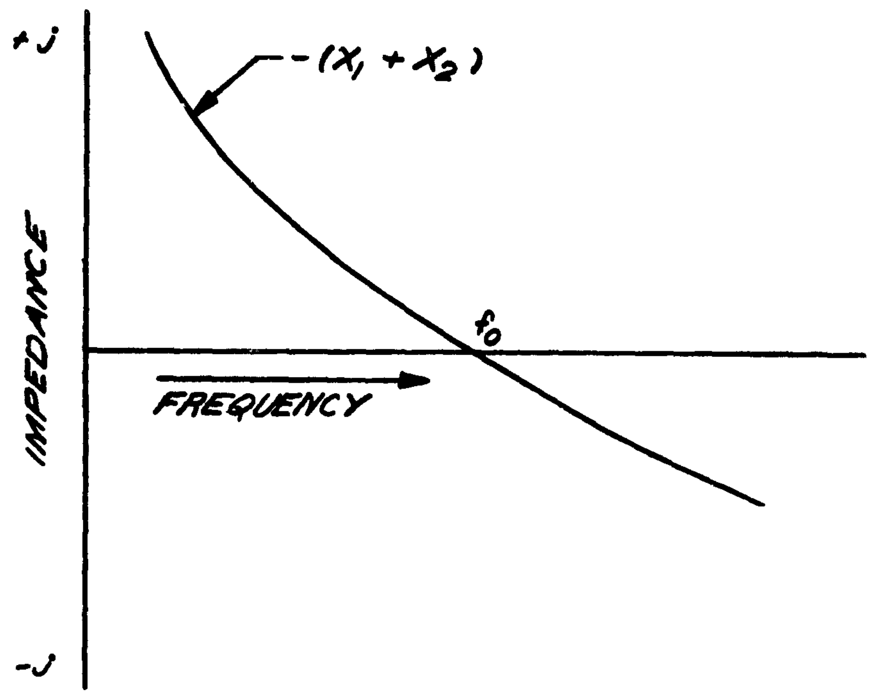


Figure 35. Plot of Equation (47)

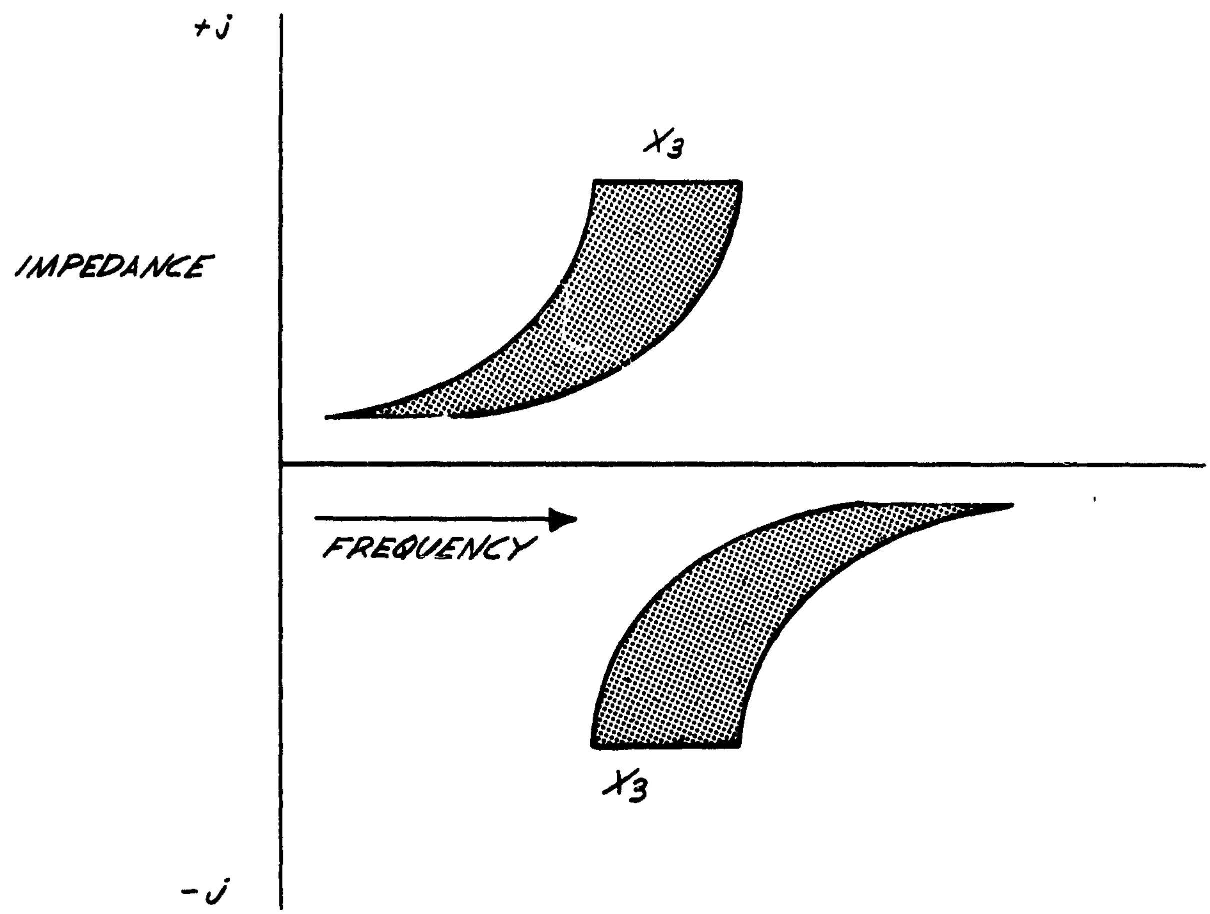


Figure 36. Effect of Variable Tuned Oscillator Plate Circuit

Figure 36 demonstrates the effect of a variable-tuned circuit which often is used in the plate circuit of an oscillator. Figure 37 is a composite of figures 35 and 36 and shows clearly the frequency range in which a given circuit could be expected to oscillate. The range between f_1 and f_2 is eliminated by the condition that the reactance of X_3 must be opposite in sign to that of X_1. Therefore, oscillation can occur only in the range f_3 to f_4. The further assumption is made that g_m is sufficiently large to sustain oscillation for the particular value of R_x used in the circuit. In figure 37, the series resonant frequency of the crystal occurs between f_2 and f_3. Thus, to operate at series resonance, the value of X_3 must be in the negative reactance shaded area.

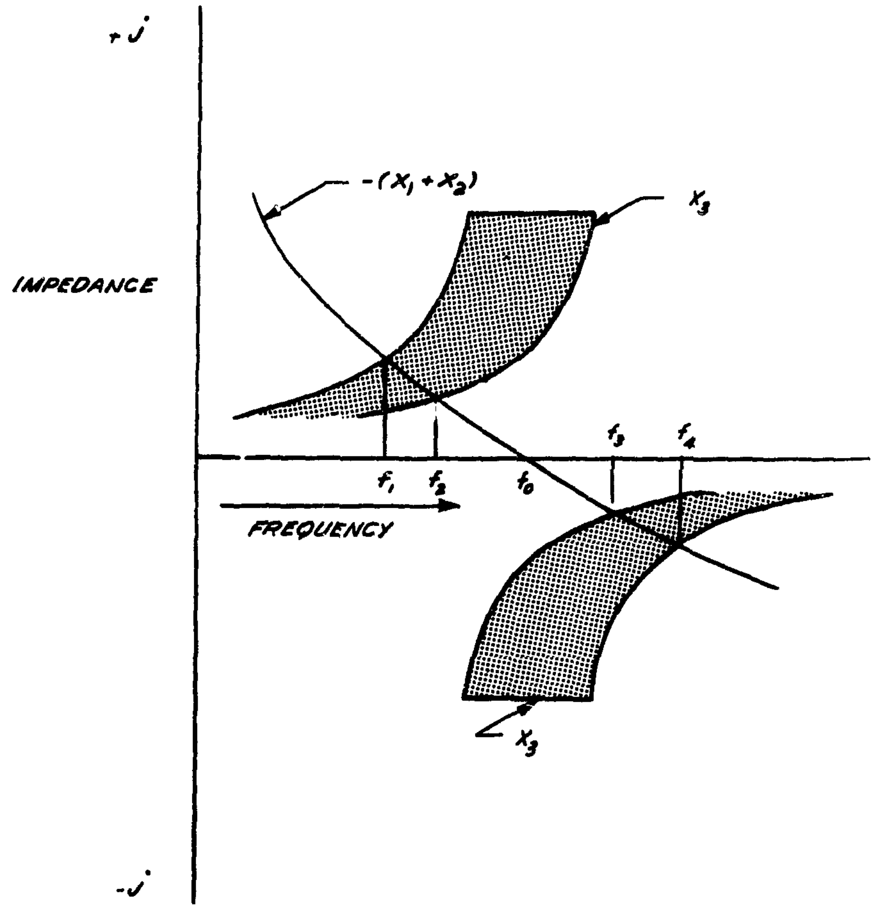


Figure 37. Combined Plot of Figures 35 and 36

The problems created by adding two sections of transmission line between the crystal and the oscillator are twofold. First, the line will add a time delay or phase shift between the plate and grid circuits. Second, the line will transform voltage, current, and impedance at the plate end (generator) to that at the load end (grid). Both of these effects are functions of the electrical length of the line and the terminating impedance.

In figure 38, if the length, L, of the sections of the line is one-half wavelength, the phase of the voltage and current will be shifted 180 degrees, but the impedance will remain unchanged. If L equals one wavelength, the voltage, current, and impedance will have the same relative phase and magnitude. Thus, for values of $L = \frac{n\lambda}{2}$, where $n = 0, 1, 2, 3$ ----, the circuit will be unaffected by the line. Therefore, the problem of

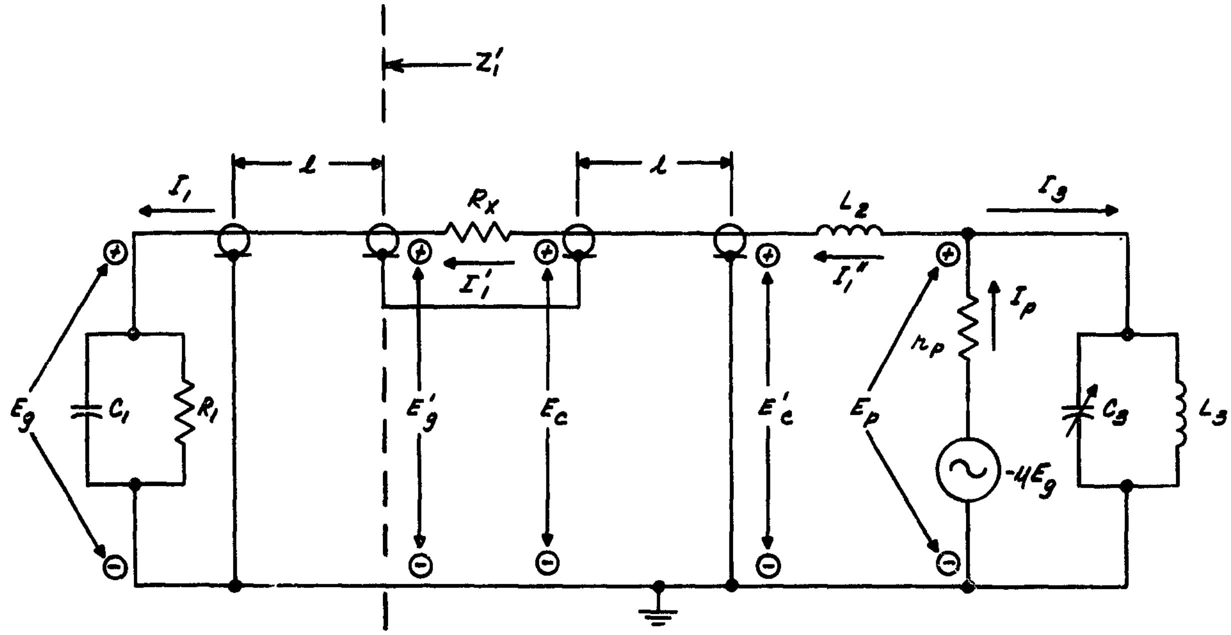


Figure 38. Transmission Line Crystal Oscillator, Equivalent Circuit

proper circuit operation arises only when the electrical length does not equal an integral half wavelength.

Referring to figure 38 and assuming that the transmission line is lossless, the equations for its performance are:

$$E_g' = E_g \left(\cos BL + j\,\frac{Z_o}{Z_1}\sin BL\right)$$

$$E_c' = E_c \left(\cos BL + j\,\frac{Z_o}{R_x + Z_1}\sin BL\right)$$

$$I_1' = I_1 \left(\cos BL + j\,\frac{Z_1}{Z_o}\sin BL\right)$$

$$I_1'' = I_1' \left(\cos BL + j\,\frac{R_x + Z_1}{Z_o}\sin BL\right)$$

where:

Z_o = characteristic impedance of the transmission line

Z_1' = impedance seen from receiving end of the transmission line

Z_1 = impedance seen from generator end of the transmission line

B = velocity factor in electrical degrees/unit length

L = physical length of line.

From these equations, Z_1' may be calculated from E_g and I_1. It should be noted that the phase shift of the line is a function of the load impedance. Therefore, the

two sections of the line will have different phase shifts because of the crystal resistance, R_x, added between them. In the same manner, the schematic location of L_2 also affects the total phase shift between the grid and the plate of the oscillator. To indicate the degree of the effect of the crystal resistance at the center of the transmission line, a set of calculations was made at 92 mc using coaxial line 0.6 wavelength long on each side of the crystal and L_2 connected as shown in figure 38. These calculations resulted in the following data:

For a 15-ohm crystal resistance,

$$E_c' = E_c\,(1.36\ \underline{|189.1^\circ})$$

$$I_1'' = I_1'\,(1.46\ \underline{|190^\circ})$$

For a 60-ohm crystal resistance,

$$E_c' = E_c\,(0.68\ \underline{|202^\circ})$$

$$I_1'' = I_1'\,(1.63\ \underline{|207.8^\circ})$$

Thus, not only is the phase shift changed by the value of the crystal resistance but also the resistance and reactance as viewed from the plate end of the transmission line.

At any particular frequency, the test oscillator will be more likely to operate satisfactorily over a wide range of crystal resistances if the length of the line is adjusted to an integral number of one-half wavelengths. It is recommended that if difficulty is experienced at a particular frequency a special section of line be assembled to provide $n\frac{\lambda}{2}$ wavelengths.

4.2.4 Audio Amplifier. The audio amplifier circuit performs three functions:

(1) Provides a means of a-c coupling to the high impedance output of the phase detector.

(2) Amplifies the a-f output signal from the phase detector by a factor of ten.

(3) Limits the bandwidth of the system from 20 cps to 4000 cps.

The various RC time constants in the amplifier have been adjusted to provide 1-db roll-off at 20 cps. The 4-kc low-pass filter provides the attenuation at the high end of the audio range. The 4-kc upper limit to the audio pass band was chosen to allow the second harmonic of the maximum vibration frequency to appear on the readout oscilloscope. Past

experience has shown that some crystal mounts exhibit nonlinear characteristics and produce a definite second harmonic component in the phase-deviation output. The over-all response curve of the audio amplifier is shown in figure 39.

A schematic diagram of the audio amplifier is shown in figure 40. The amplifier uses both sections of a 12AX7 tube. A very large amount of cathode degeneration has been designed into each stage to provide a high degree of gain stability. The potentiometer between the two stages allows setting of the over-all amplifier gain at ten. This factor of ten was dictated by the maximum oscilloscope sensitivity (10 millivolts/cm) and the maximum design resolution of the phase detector (1 millivolt). The gain control is located on the main chassis under the dust cover as shown in figure 6.

The output impedance of the amplifier is 10K ohms and is determined by the characteristic impedance of the low-pass filter.

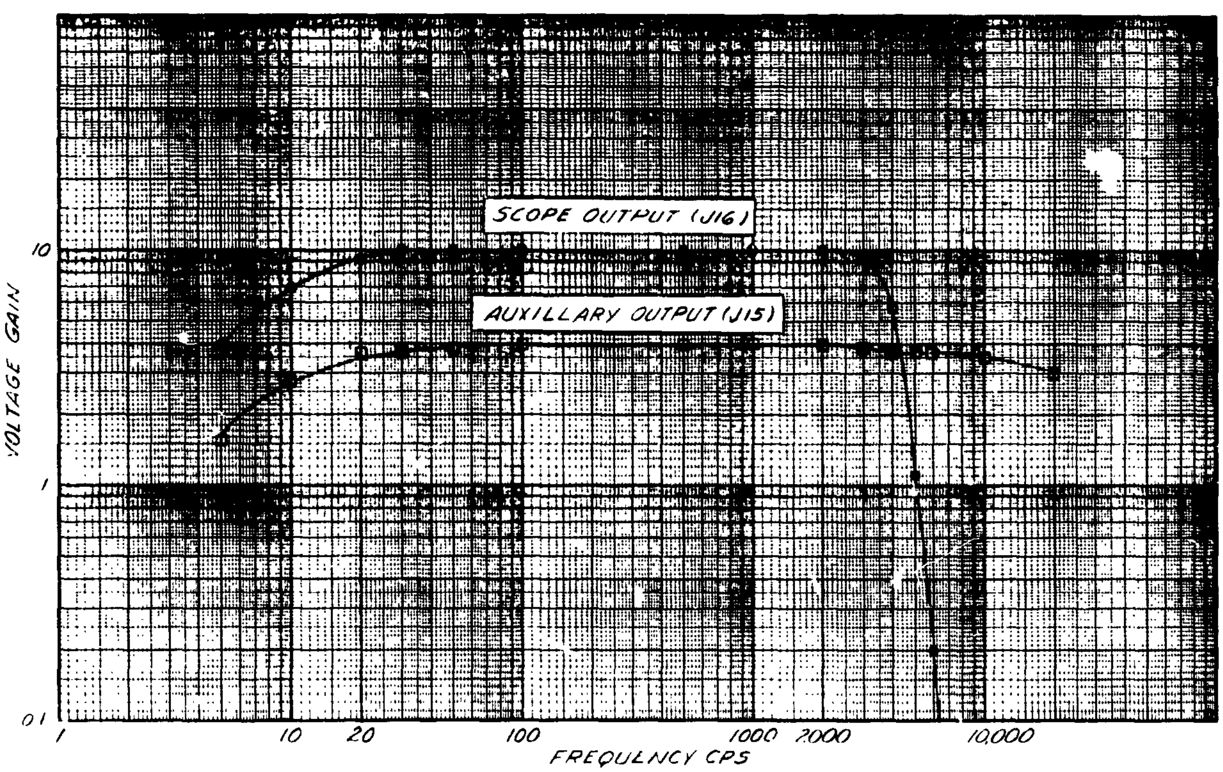


Figure 39. Audio Amplifier Frequency Response

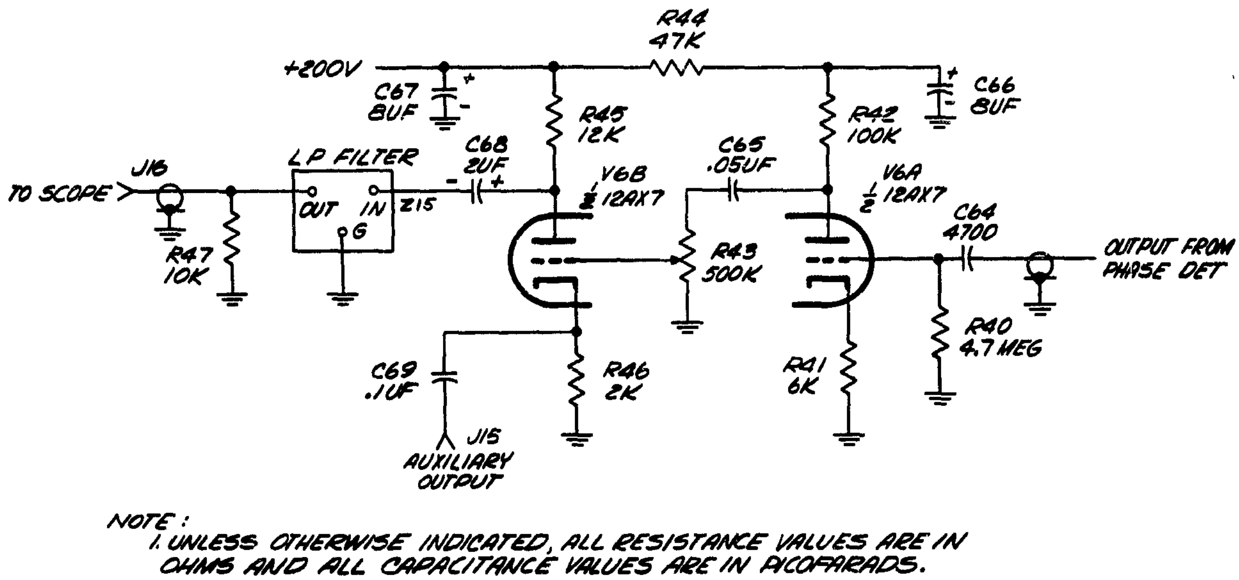


Figure 40. Audio Amplifier, Schematic Diagram

The audio amplifier is in the circuit between the output of the phase detector and the oscilloscope only when the function switch is on the operate position. In the calibrate and LOCK positions, the amplifier is removed from the circuit.

The auxiliary output of the amplifier is located at a BNC jack on the rear of the main chassis as shown in figure 5. This output is provided in order that external differentiator network may be connected to the equipment to allow measurement of the frequency deviation of the crystal under test. The output impedance as a function of audio frequency is shown in figure 41. Obviously, the slope of this curve may be altered by changing the value of C69.

4.2.5 Integrating Network. The integrating network was designed in accordance with the material presented in paragraph 4.3 of this report. A schematic of the network and associated components is shown in figure 42.

The +1 volt d-c applied to the integrating capacitor is used to maintain a small charge on the capacitor. The output of the phase detector will increase or decrease this charge, depending upon the voltage at which a locked-loop condition can be achieved.

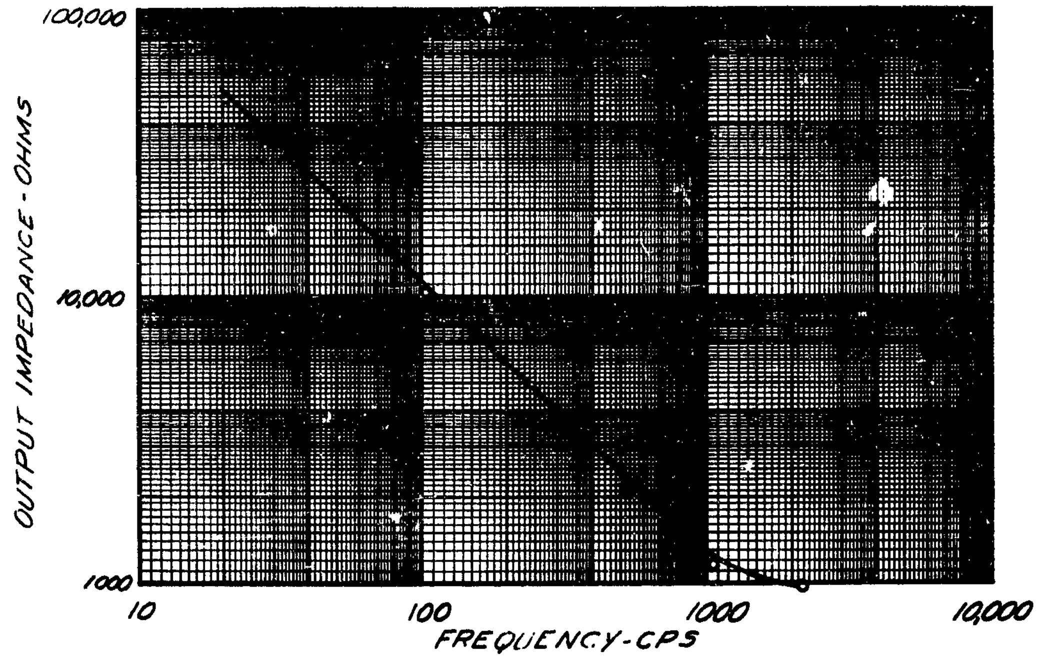


Figure 41. Output Impedance of Auxiliary Output J15

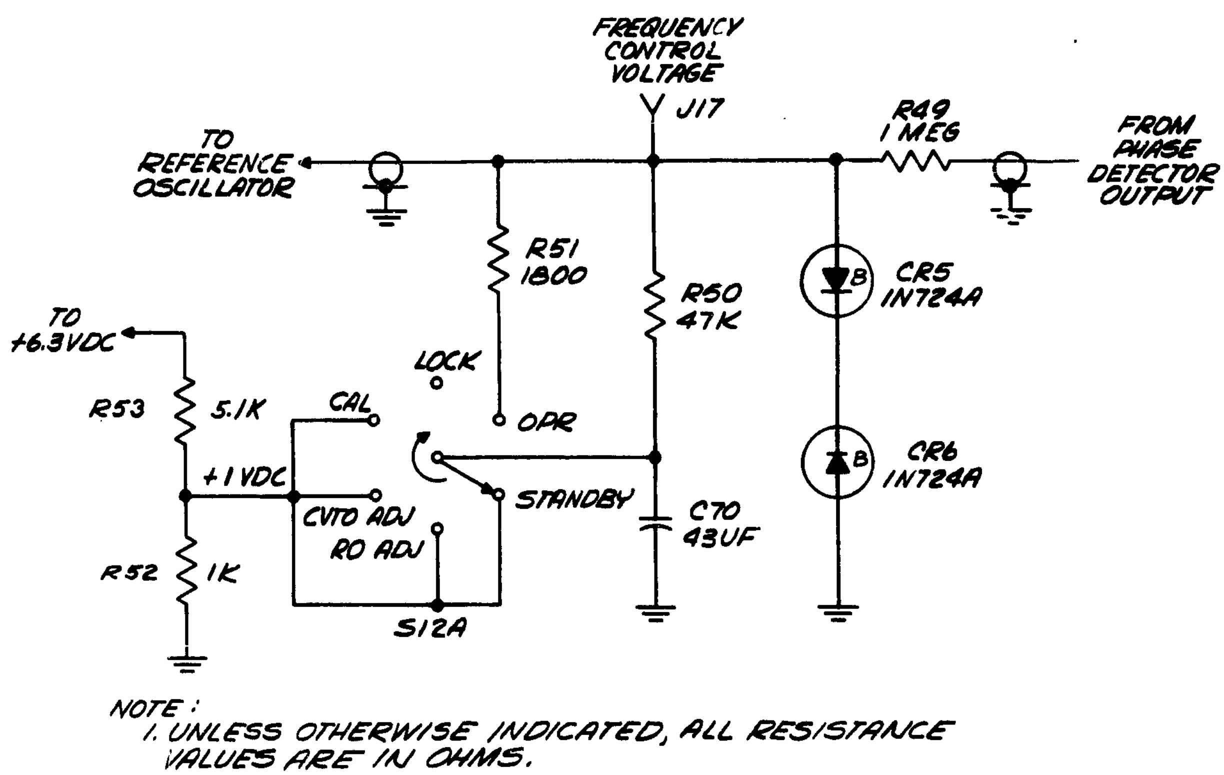


Figure 42. Integrating Network, Schematic Diagram

The voltage also provides a constant level on the reference oscillator voltage variable capacitors in the RO ADJ, CVTO ADJ, and CAL positions of the function switch. The 1N724A Zener diodes have 27-volt breakdown points and thus prevent the voltage applied to the voltage-variable capacitor from exceeding this value. The back-to-back arrangement of the diodes clips both positive and negative half cycles of the frequency control voltage.

4.2.6 Power Supplies. The equipment specification for this contract required the use of d-c power supplies for not only the high voltage but also for the filament or heater supplies. It further required that the equipment be operable from either 115-volt or 230-volt, 60-cycle supply line.

The equipment performance is dependent largely upon the ability of the power supplies to furnish a very pure d-c voltage for both the plate voltages and the heater voltages. In the initial equipment calculations, it was decided to design for an unlocked-loop output voltage of 180 volts peak-to-peak for 180 electrical degrees at 1 mc. This would mean that a 1-millivolt output signal from the phase-locked loop would be equivalent to a peak-to-peak phase deviation of 1 millidegree. This would provide for a safety factor of six from the specifications. In order to achieve such performance, the plate supply voltage would have to possess less than 1 millivolt of ripple voltage.

The high-voltage supply used in this equipment was purchased as a commercial unit from ACDC Electronics, Burbank, California. The Model RPM 200-200 is used to supply the 200 volts d-c at 200 milliamperes. This power supply operates from 105 to 125 volts a-c, 50 to 400 cps supply line and is regulated electronically to 0.05 percent for no-load to full load or, for line changes of ±10 percent, under any conditions of load current of line-voltage variations within the limits specified. The output ripple voltage is 1 millivolt rms maximum under any conditions of load current or line voltage within the limits specified. This power supply is constructed in a one-piece cast-aluminum housing and is the right-hand unit on the power supply chassis shown in figure 43. The high-voltage supply furnishes the +200 volts d-c to the various vacuum tube circuits of the equipment. Tests

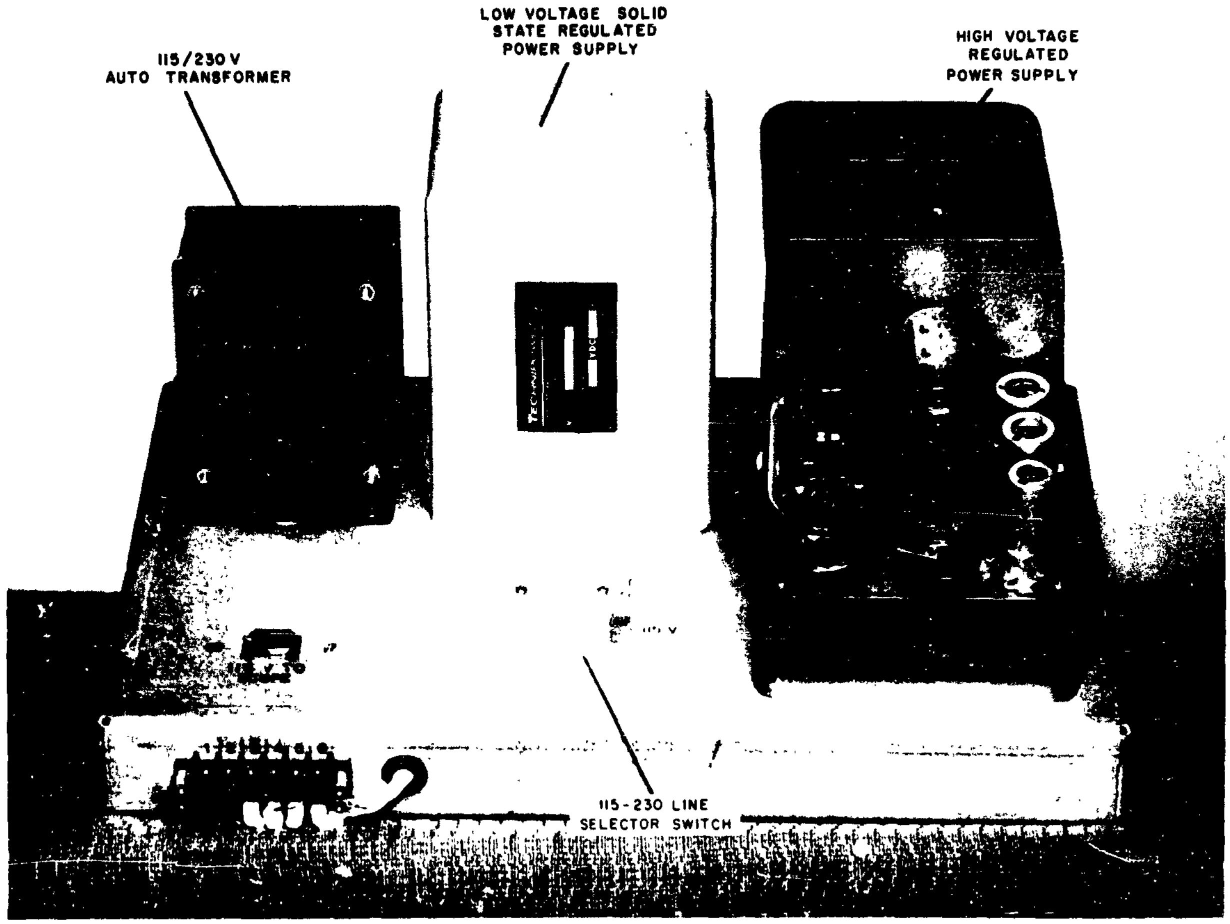


Figure 43. Power Supply Chassis

have been conducted on this unit and show the ripple and regulation to be well within the manufacturer's specifications.

Inasmuch as any stray voltages or fields introduced into the main chassis would detract from the ultimate resolution of the equipment, the heaters of the vacuum tubes in the equipment all are operated on 6.3 volts d-c, and no 60-cycle power wiring appears inside of the r-f or measurement circuitry.

The heater supply for the equipment is a miniature solid-state regulated power supply manufactured by Technipower, Incorporated, South Norwalk, Connecticut. The supply is regulated ±0.05 percent and has a ripple voltage of less than 1 millivolt rms

for specified variations in line and load conditions. The output voltage of this unit may be changed by means of an internal adjustment, access to which is through a small hole on the under side of the power supply chassis. Under full load conditions, the output is set for 6.3 volts d-c at the audio amplifier tube socket. Proper setting of this control is necessary, as the General Electric ceramic tubes are somewhat sensitive to improper heater voltage. If the heater voltage is too low, the gain of the various circuits will be reduced. If the heater voltage is too high, the tube life will be considerably shortened. General Electric recommends maintaining the heater voltage at 6.3 volts, ±5 percent. This power supply is more than adequate in this application. This unit is the center subassembly on the power supply chassis shown in figure 43.

Also included on the power supply chassis is an autotransformer which may be switched into operation if it is desired to operate the equipment from a 230-volt, 60-cycle supply source. This transformer is capable of handling both the equipment power supplies and the oscilloscope. In order to operate from a 230-volt line, the aluminum-switch cover block must be removed, the switch thrown to the 230-volt position, and the cover block replaced to indicate 230-volt operation is in effect.

The ON-OFF switch and the line fuses for the equipment are mounted on the power control panel on the front of the cabinet turret.

4.2.7 Oscilloscope. This equipment uses a Hewlett-Packard Model 120/AR oscilloscope for the readout. This oscilloscope has more than adequate frequency response (200 kc) and is the most economical unit commercially available which had adequate sensitivity (10 millivolts/cm).

This oscilloscope has a calibrator circuit whose operation allows compliance with the technical requirement for adjusting the phase-deviation pattern to the center 50 percent of the oscilloscope graticule.

In order to protect the special circuitry in the audio amplifier of the phase stability analyzer, the vertical input terminals to the oscilloscope have been removed from the front panel and a BNC type coaxial connector placed on the rear of the oscilloscope

chassis. This connector is connected directly to the scope output jack on the rear of the main chassis of the phase-stability analyzer as shown in figure 3.

4.2 Cabinet. The r-f phase stability analyzer is mounted in a semicustom modular enclosure system manufactured by Amco Engineering Company, Chicago, Illinois. This cabinet is shown in figures 2 and 3. It is mounted on casters and finished with a gray vinyl scuff-resistant paint. The side panels and other sections which have sounding board tendencies have had styrofoam sheets cemented to their inside surfaces. This provides sufficient dampening so that these sections will not amplify any vibration due to acoustical or mechanical pickup. Such construction was necessitated because of the microphonic sensitivity of the main chassis. This does not eliminate the undesirable effects due to acoustical pickup completely, but it does reduce the effects to a low level. A set of tools for equipment disassembly are mounted inside the cabinet as shown in figure 3.

The drawer at the bottom front of the cabinet is available for storage of the various crystal mounting parts, coaxial cable sections, and the crystal vibration-test oscillator assembly as shown in figure 44. The coaxial cable sections for adjusting the length of the transmission line between the crystal mounting block and the test oscillator are stored in a recess directly under the test oscillator case.

Due to the overhang of the turret, the power supply chassis was mounted low and to the rear of the cabinet. This arrangement provides mechanical stability to the complete equipment when it is being moved.

4.3 Mathematical Analysis of Complete System. Let the phase-lock loop be represented by the block diagram of figure 45. For purposes of this analysis, each of the blocks will be represented as follows:

4.3.1 Crystal-Vibration-Test Oscillator. As shown in figure 46, (ω_1) is the average angular frequency of the CVTO and is defined to be a constant. For purposes of this analysis, $N_1 = \Delta\omega_1$, the frequency modulation caused by vibration of the crystal. Note that $\int_0^t \Delta\omega_1 \, dt = \Delta\phi_1$, the phase change of the CVTO caused by vibration. For Laplace transforms, $\frac{\Delta\omega_1}{S}$ in the S plane is equivalent to $\int_0^t \Delta\omega_1 \, dt$ in the time domain.

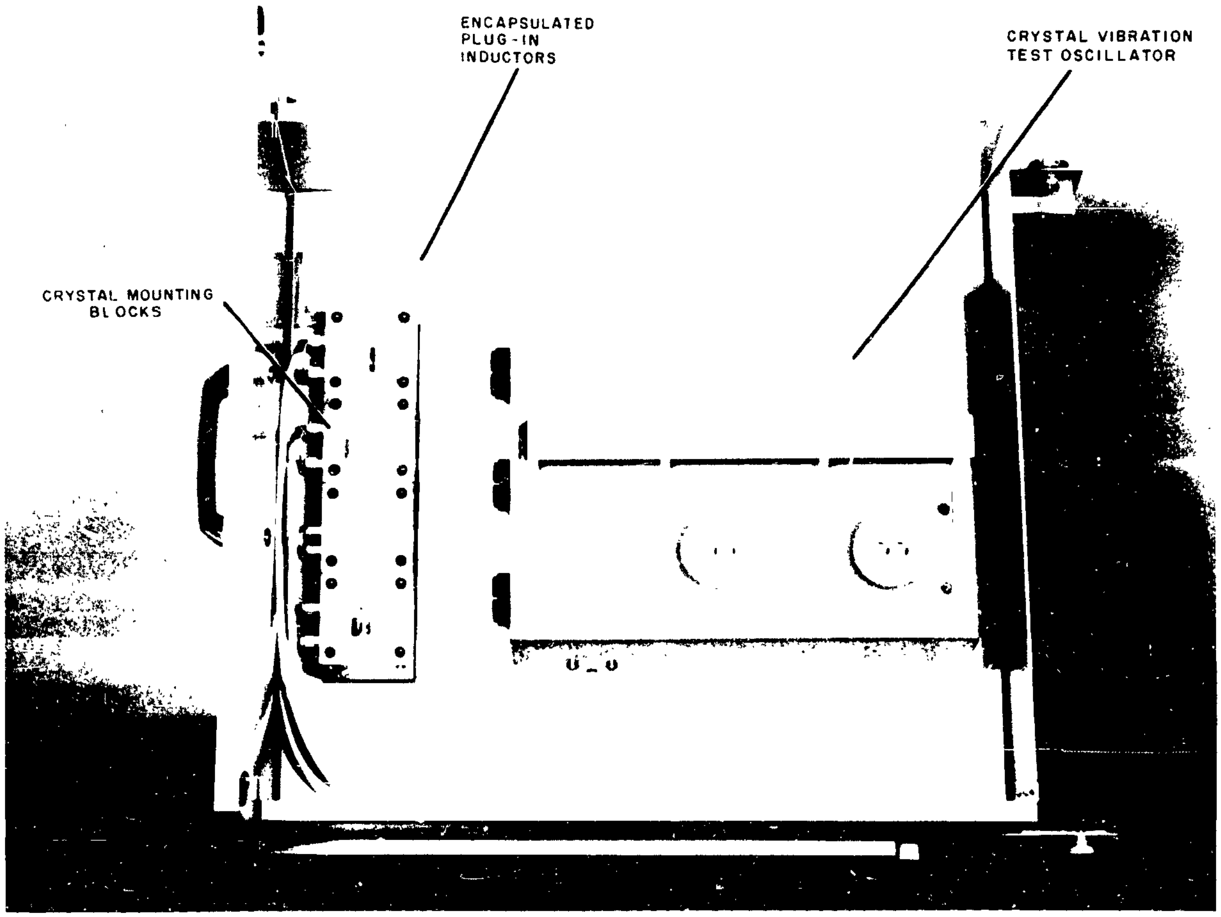


Figure 44. Cabinet Drawer, Interior View

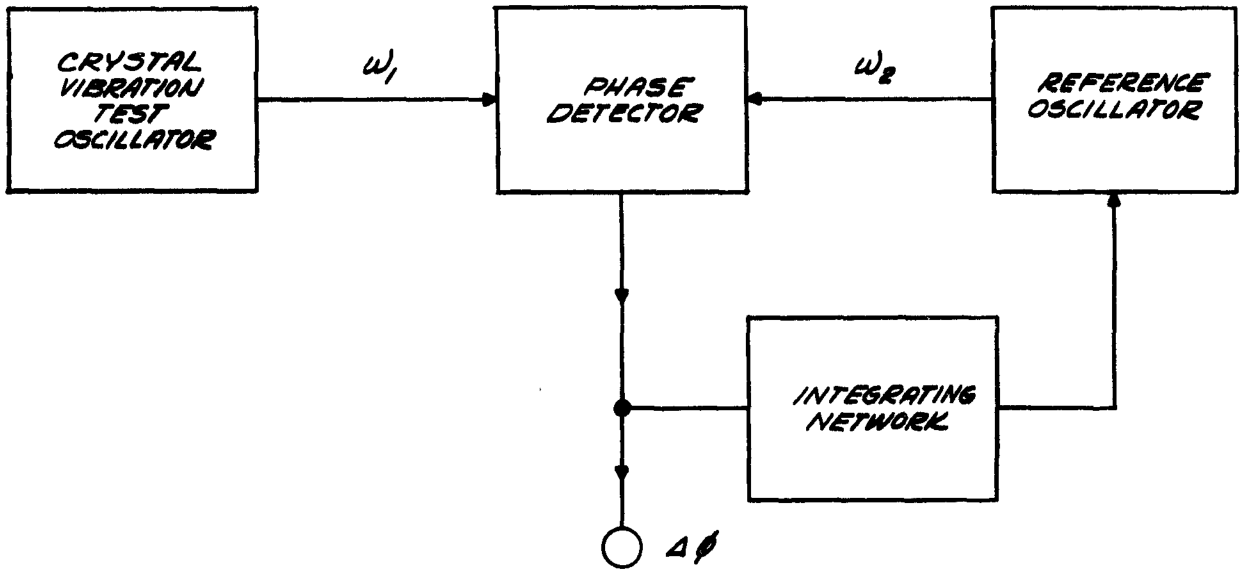


Figure 45. Phase-Lock Loop, Basic Block Diagram

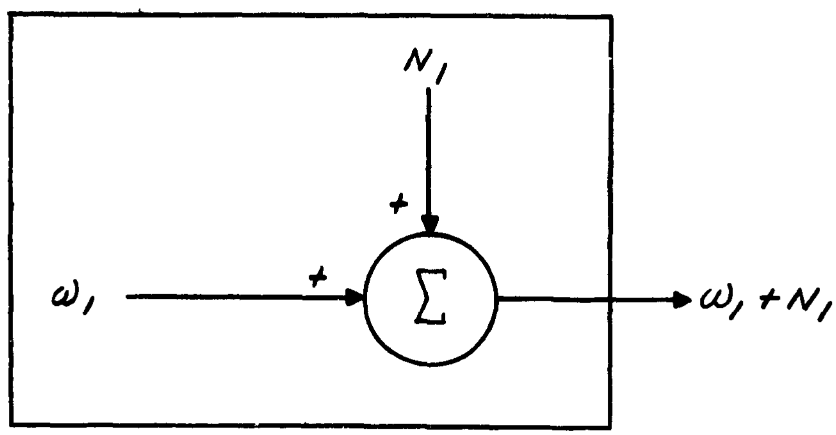


Figure 46. Mathematical Equivalent of Crystal-Vibration-Test Oscillator

4.3.2 Reference Oscillator. As shown in figure 47, Ω is the angular frequency of the reference oscillator with zero voltage on the control element. K_2 is in radians/sec/volt. N_2 is any frequency modulation on the output of the reference oscillator other than that caused by the control voltage. Since N_2 is considerably less than the modulation of the CVTO, it will be neglected in this analysis.

4.3.3 Phase Detector. As shown in figure 48, K_1 is the sensitivity of the phase detector in volts/radian. The $\frac{1}{S}$ factor represents integration in the S plane.

4.3.4 Integrating Network. As shown in figure 49, F(S) is the transfer function of the integrating network in the S plane.

Replacing the blocks of figure 45 by their mathematical equivalents yields the diagram of figure 50. Figure 51 shows figure 50 redrawn to appear as a conventional servo system.

From figures 51 and 52 it can be seen that:

$$G(S) = \left(\frac{1}{S}\right)(K_1) = \frac{K_1}{S} \tag{49}$$

and

$$H(S) = F(S)\,K_2$$

According to conventional feedback theory

$$C = \frac{G(S)\,R}{1 + H(S)\,G(S)} = E_0 \tag{50}$$

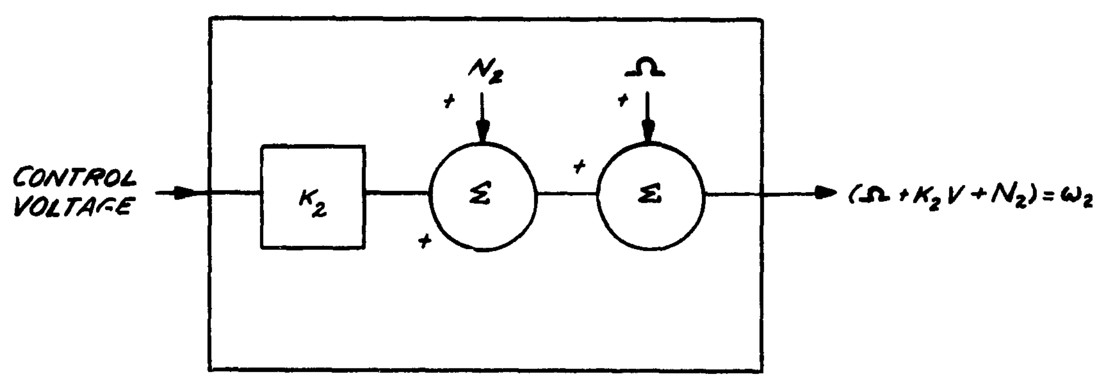


Figure 47. Mathematical Equivalent of Reference Oscillator

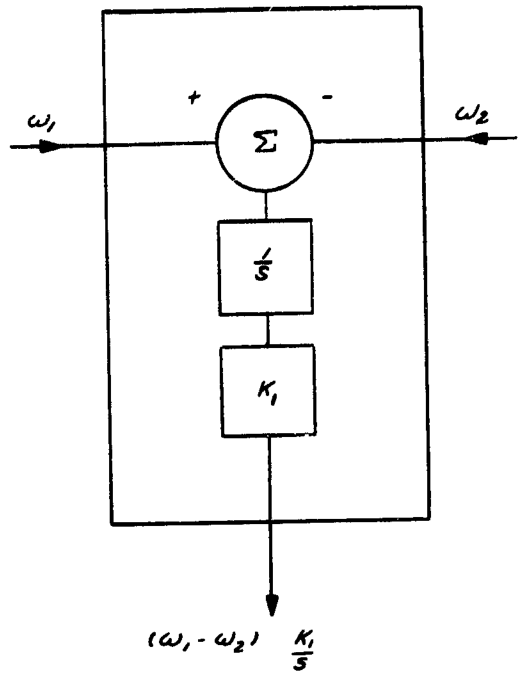


Figure 48. Mathematical Equivalent of Phase Detector

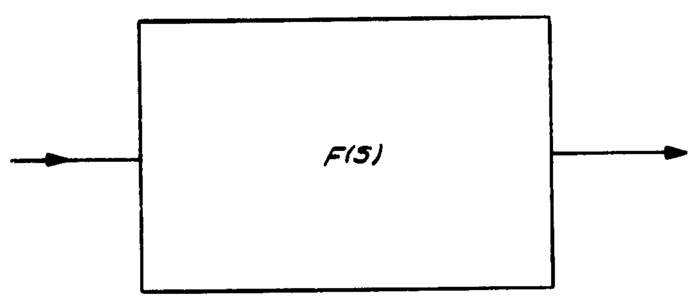


Figure 49. Mathematical Equivalent of Integrating Network

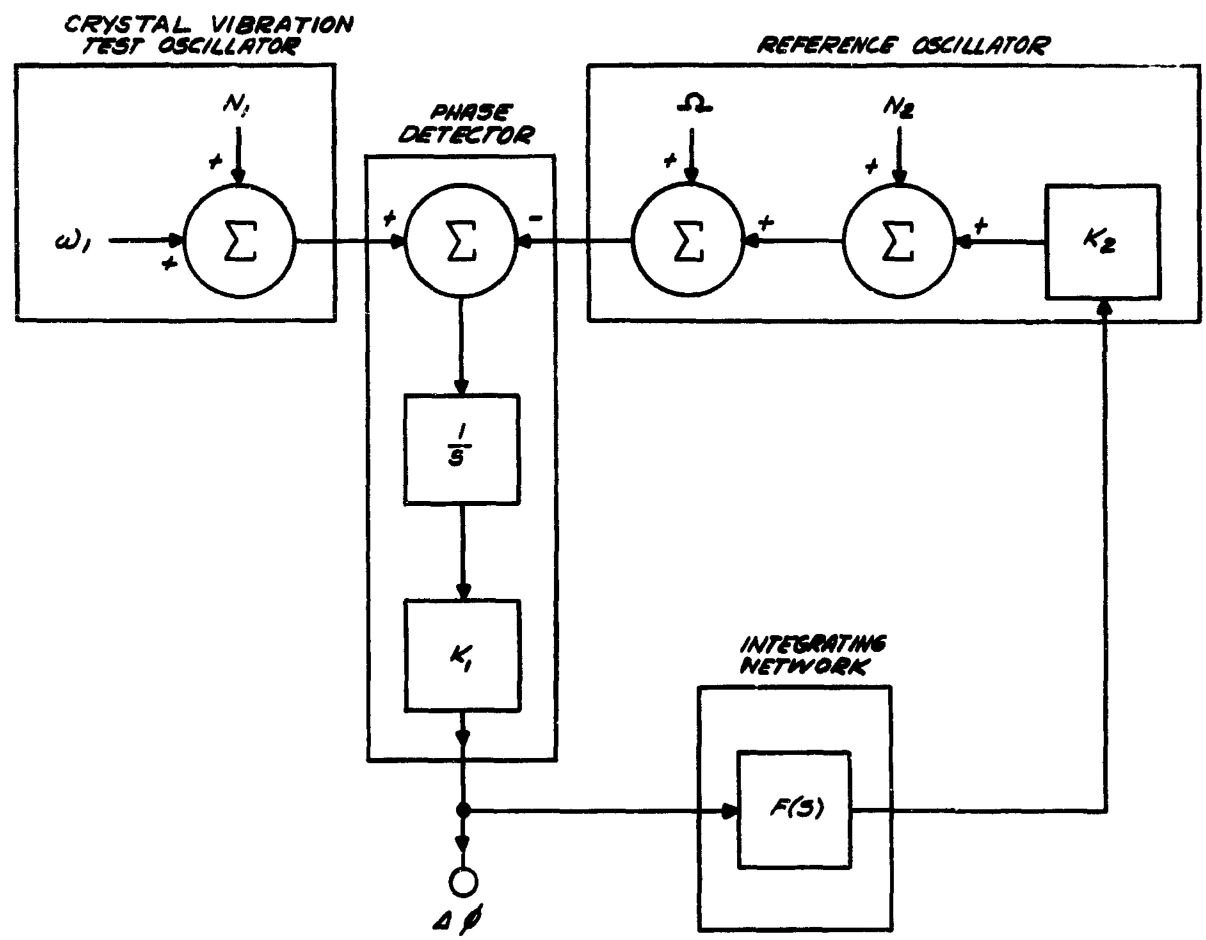


Figure 50. Mathematical Equivalent of Phase-Lock Loop, Block Diagram

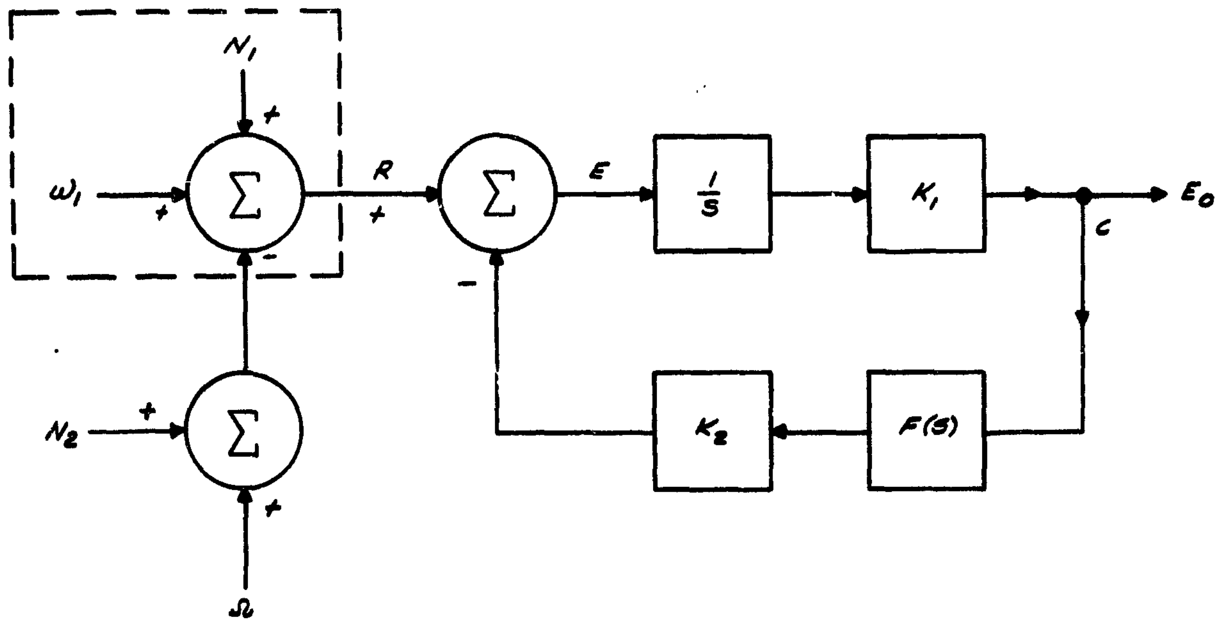


Figure 51. Rearrangement of Mathematical Equivalent of Phase-Lock Loop, Block Diagram

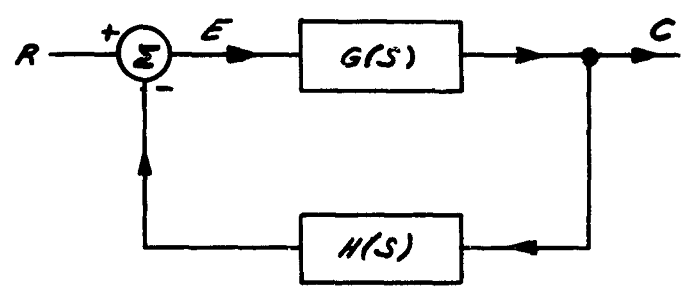


Figure 52. Simplified Phase-Lock Loop, Block Diagram

but $R = \omega_1 + N_1 - N_2 - \Omega$. Substitution in equation (50) yields:

$$C = \frac{G(S)[\omega_1 + N_1 - N_2 - \Omega]}{1 + H(S)G(S)} = E_0$$

It is assumed that N_2 is negligible, $\omega_1 = \Omega, N_1 = \Delta\omega_1, C = E_0$.

$$E_0 = \frac{G(S)\left[\omega_1 + \Delta\omega_1 - 0 - \omega_1\right]}{1 + H(S)G(S)}$$

$$E_0 = \frac{G(S)\Delta\omega_1}{1 + H(S)G(S)}$$

but $G(S) = \frac{K_1}{S}$

From equation (49), therefore,

$$E_0 = \frac{K_1\left(\frac{\Delta\omega_1}{S}\right)}{1 + H(S)\,G(S)} \tag{51}$$

However, in the S plane $\frac{\Delta\omega_1}{S}$ is equivalent to $\int_0^t \Delta\omega_1 \, dt$ in the time domain

and therefore $\frac{\Delta\omega_1}{S} = \Delta\phi_1$

$$E_0 = \frac{\Delta\phi_1 K_1}{1 + H(S)G(S)}$$

$$\left(\frac{E_0}{\Delta\phi_1}\right) = \frac{K_1}{1 + H(S)G(S)} = \frac{K_1}{1 + \frac{K_1 K_2 F(S)}{S}} \tag{52}$$

where $H(S)\,G(S) = \frac{K_1 K_2 F(S)}{S}$

The integrating network F(S) may be varied, and a number of networks can be used with acceptable results. The conventional lag network shown in figure 53 is used with this equipment, since it reduces the capture range less than most other simple networks for a given bandwidth.

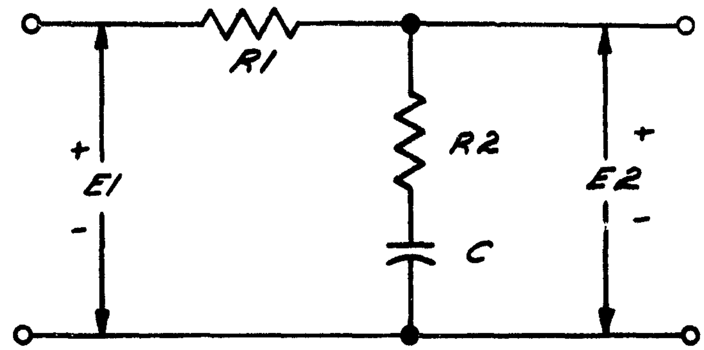


Figure 53. Simplified Integrating Network, Schematic Diagram

It can be shown[1] that for the network of figure 53

$$\frac{E_2(S)}{E_1(S)} = F(S) = \frac{1 + T_2 S}{1 + T_1 S} = \frac{T_2\left(S + \frac{1}{T_2}\right)}{T_1\left(S + \frac{1}{T_1}\right)} \tag{53}$$

where $T_1 = (R_1 + R_2)C$ and $T_2 = R_2 C$.

Since $\frac{E_o}{\Delta\phi_i} = \frac{K_1}{1 + H(S)G(S)}$ and represents the gain of the system insofar as the quantity of interest is concerned, it is desirable to make this ratio as large as possible as well as constant in the frequency range from 20 to 2000 cps (125.6 to 12,560 rad/sec). Therefore, K_1 is made as large as possible, and H(S)G(S) is made negligible compared to 1.

Then $\left(\frac{E_o}{\Delta\phi_i}\right) \doteq K_1$.

Since the product H(S)G(S) decreases with frequency, if $0.1 \geq |H(S)G(S)|$ at 20 cps then $\frac{E_o}{\Delta\phi_i} \doteq K_1$ over the entire frequency range of interest.

[1]Reference 1 of Bibliography, p 107.

If no integrating network were used, $F(S) = 1$, and the capture range* would be a maximum and equal the hold range** which $= \pi K_1 K_2$.

If the integrating network is used, the capture range is approximately equal to $\pi K_1 K_2 \sqrt{\frac{2T_2}{T_1}}$; however, the hold range is unchanged. In this manner the gain of the system can be reduced in the low frequency region (around 20 cps)[2] without reducing the hold range. If, on the other hand, K_2 were reduced to decrease the gain, the hold range and capture range would be reduced proportionally.

For the design of F(S), then

$$0.1 = |G(S)\,H(S)| \text{ at 20 cps}$$

$$0.1 = \left|\frac{K_1 K_2 F(S)}{S}\right| \text{ and } F(S) = \frac{T_2}{T_1}\frac{\left[S + \frac{1}{T_2}\right]}{\left[S + \frac{1}{T_1}\right]}$$

Therefore,

$$0.1 = \left|\left(\frac{K_1 K_2}{S}\right)\left(\frac{T_2}{T_1}\right)\frac{\left[S + \frac{1}{T_2}\right]}{\left[S + \frac{1}{T_1}\right]}\right|_{\text{at 20 cps.}} \tag{54}$$

For satisfactory transient response, it is necessary that $\frac{1}{T_2} < \omega_c$, where ω_c is the angular frequency[3] at which $H(S)G(S) = 1$, since the slope of $H(S)G(S) = 20$ db/decade between f_c and 20 cps (as shown in figure 54), and since $H(S)G(S) = 20$ db at 20 cps, f_c occurs at 2 cps. Therefore, $\frac{1}{T_2}$ (maximum) $= 12.56$ rad/sec $= 2$ cps. T_2 (minimum) $= 0.0796$ seconds.

---

*Capture range is defined to be the maximum difference in frequency between the two oscillators for which lock will occur when the loop is closed.

**Hold range is defined to be the maximum frequency difference between unlocked oscillators for which lock can be maintained.

[2, 3] References 2 and 3 of Bibliography, p 107.

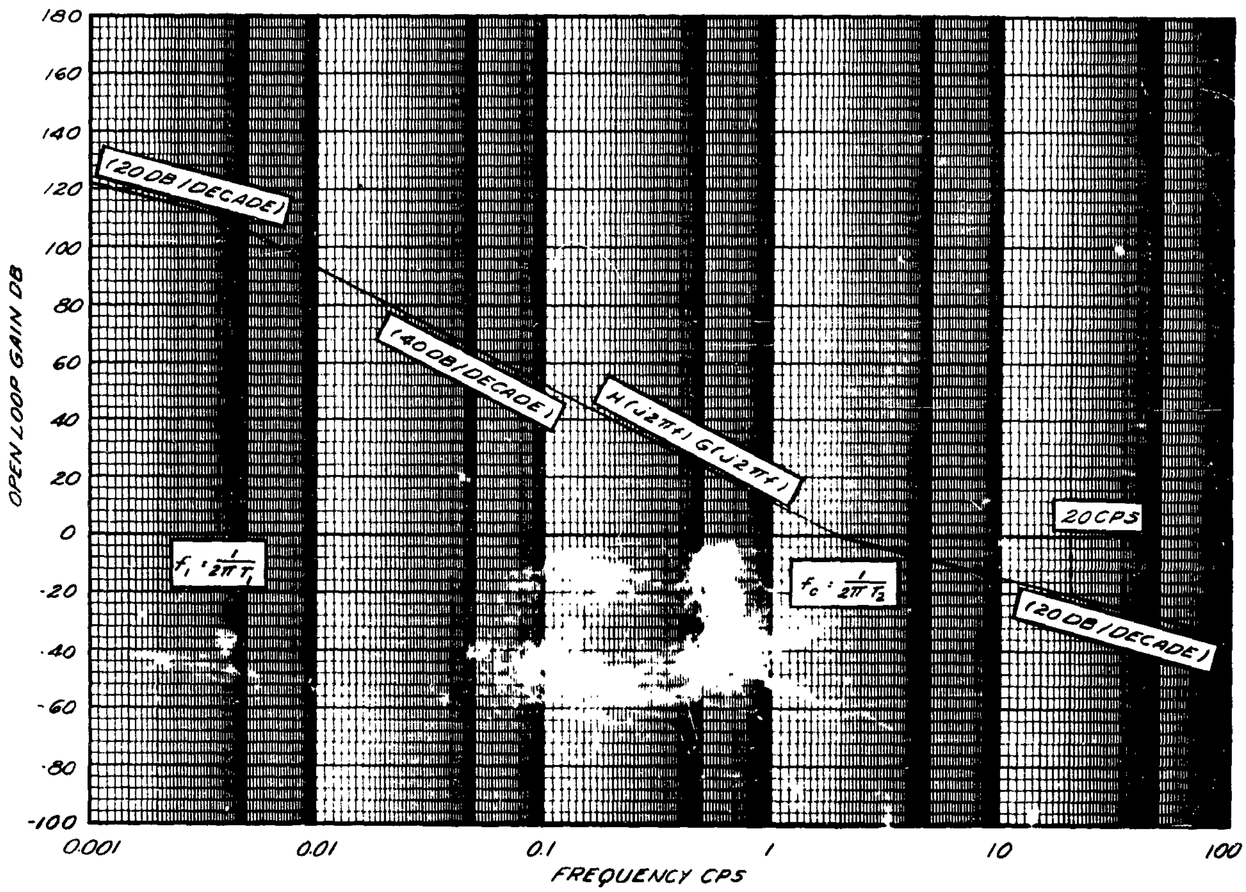


Figure 54. Idealized Open Loop Gain Curve

Then putting 0.0796 into equation (54) for T_2 yields

$$0.1 = \left| \left(\frac{K_1 K_2}{j\,2\pi(20)} \right) \left(\frac{0.0796}{T_1} \right) \left(\frac{j2\pi 20 + 12.56}{j2\pi 20 + \frac{1}{T_1}} \right) \right|$$

$$0.1 \approx \frac{K_1 K_2 0.0796}{125.6\ T_1}$$

$$T_1 = 0.00633 K_1 K_2$$

The capture range of the system is therefore $= \pi K_1 K_2 \sqrt{\dfrac{2T_2}{T_1}}$

$$= \pi K_1 K_2 \sqrt{\frac{2 \times 0.0796}{0.00633 K_1 K_2}}$$

$$= 15.75 \sqrt{K_1 K_2}$$

The transient response of the system can be calculated in the following manner:

From equation (52) the response of the closed-loop system is:

$$\frac{E_o}{\Delta\phi_1} = \frac{K_1}{1 + \dfrac{K_1 K_2 F(S)}{S}} \tag{55}$$

$$\text{and, } F(S) = \frac{T_1}{T_2} \left(\frac{S + \dfrac{1}{T_2}}{S + \dfrac{1}{T_1}} \right) \tag{56}$$

Therefore,
$$\frac{E}{\Delta\phi_1} = \frac{K_1}{1 + \left(\dfrac{K_1 K_2}{S}\right)\left(\dfrac{T_2}{T_1}\right)\dfrac{\left(S + \dfrac{1}{T_2}\right)}{\left(S + \dfrac{1}{T_1}\right)}} \tag{57}$$

$$\frac{E_o(S)}{\Delta\phi_1(S)} = \frac{K_1 T_1 S \left(S + \dfrac{1}{T_1}\right)}{S^2\ T_1 + S(1 + K_2 K_2 T_2) + K_1 K_2} \tag{58}$$

$$\frac{E_o(S)}{\Delta\phi_1(S)} = \frac{K_1 S \left(S + \dfrac{1}{T_1}\right)}{S^2 + S\left(\dfrac{1 + K_1 K_2 T_2}{T_1}\right) + \dfrac{K_1 K_2}{T_1}} \tag{59}$$

Assume a unit step function input

$$\Delta\phi_1(S) = \frac{1}{S} \tag{60}$$

$$E_o(S) = \frac{K_1 \left(S + \dfrac{1}{T_1}\right)}{S^2 + S\left(\dfrac{1 + K_1 K_2 T_2}{T_1}\right) + \dfrac{K_1 K_2}{T_1}} \tag{61}$$

$$\text{Assume} \quad \left(\frac{1 + K_1 K_2 T_2}{T_1}\right)^2 < \frac{4 K_1 K_2}{T_2} \quad \text{(underdamped)} \tag{62}$$

$$E_o(S) = K_1 \frac{\left(S + \frac{1}{T_1}\right)}{\left[S + \frac{1 + K_1 K_2 T_2}{2T_1}\right]^2 + \left[\frac{K_1 K_2}{T_1} - \left(\frac{1 + K_1 K_2 T_2}{2T_1}\right)^2\right]} \tag{63}$$

$$E_o(t) = \left\{e^{-\left(\frac{1 + K_1 K_2 T_2}{2T_1}\right) t}\right\} \left\{K_1 \, \mathrm{Cos}\left[\frac{K_1 K_2}{T_1} - \left(\frac{1 + K_1 K_2 T_2}{2T_1}\right)^2\right]^{\frac{1}{2}} t\right.$$

$$\left. + K_1 \frac{\left[\frac{1}{T_1} - \frac{1 + K_1 K_2 T_2}{2T_1}\right]}{\left[\frac{K_1 K_2}{T_1} - \left(\frac{1 + K_1 K_2 T_2}{2T_1}\right)^2\right]^{\frac{1}{2}}} \mathrm{Sin}\left[\frac{K_1 K_2}{T_1} - \left(\frac{1 + K_1 K_2 T_2}{2T_1}\right)^2\right]^{\frac{1}{2}} t\right\} \tag{64}$$

Since the characteristics of both the phase detector and the reference oscillator change with frequency, it is impossible to maintain the idealized curve of figure 54 over the entire range from 1 to 110 mc. In practice, the integrating network has been designed as a compromise over the frequency range so that for 1-mc operation f_c occurs below 2 cps and for 110-mc operation f_c occurs above 2 cps. The result is that the system is somewhat underdamped at 1 mc and overdamped at 110 mc. Figure 55 shows the resulting closed-loop gain of the servo loop at 1 mc and also at 110 mc. Figure 56 shows the corresponding phase shift.

Figure 57 shows the frequency response of the entire system including the audio amplifier.

4.4 System Operation and Performance.

4.4.1 System Operation. Operation of the r-f phase stability analyzer is reasonably simple and straightforward once the over-all system and the idiosyncrasies of the particular unit are understood. As might be expected, adjustment and tuning are more critical in the vhf region, but with proper attention to details, no difficulties should be experienced even at 110 mc.

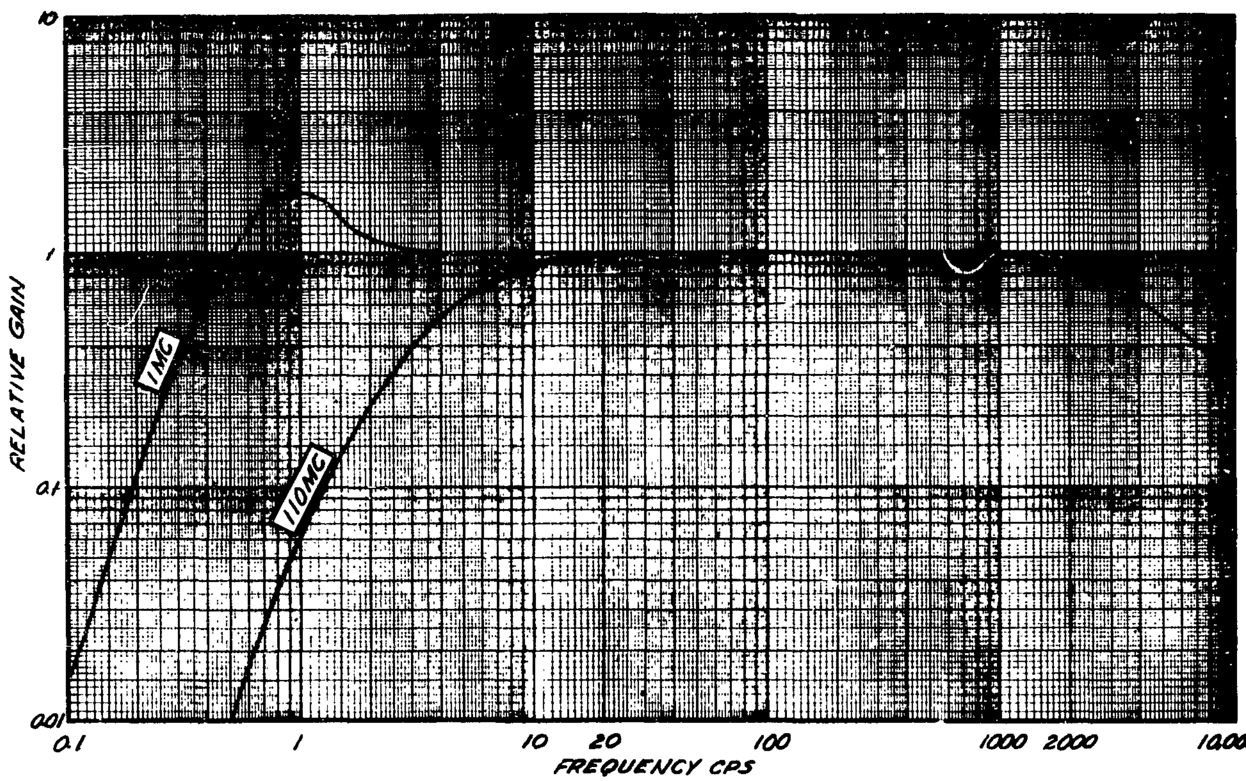


Figure 55. Gain of Phase-Locked Loop

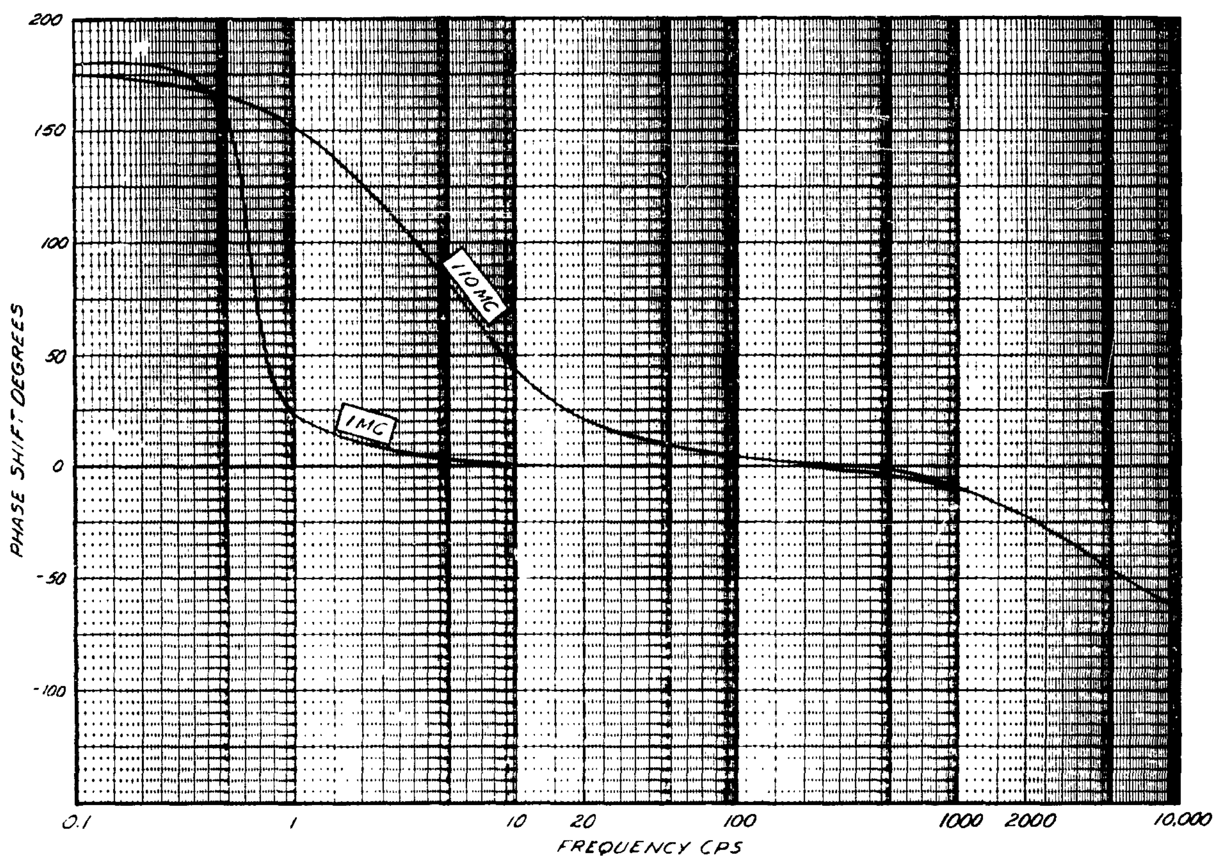


Figure 56. Phase Shift of Phase-Locked Loop

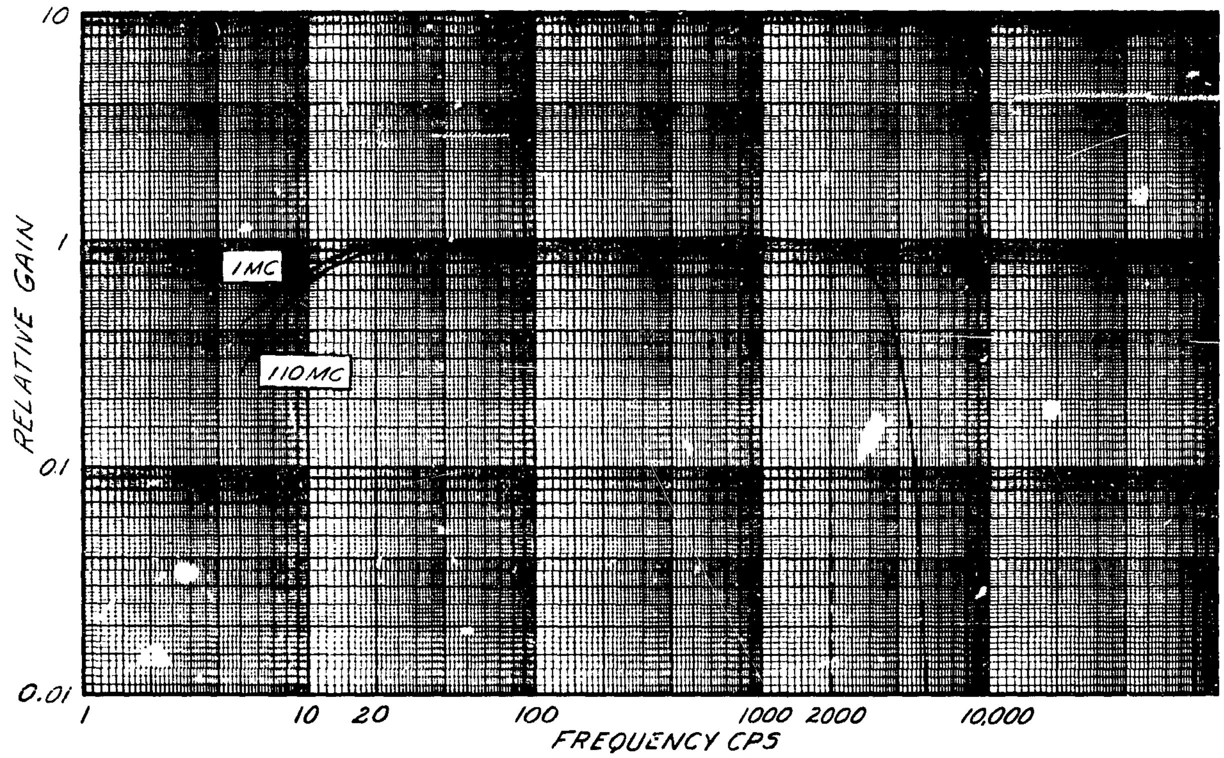


Figure 57. System Gain Curve

The following material also is presented in the instruction manual for this equipment.

4.4.1.1 Preliminary Assembly.

a. Remove the crystal-vibration test-oscillator from its drawer. Remove the 15-foot length of coaxial lead found under the CVTO.

b. Set the CVTO on a stable footing near the vibration table. Set the main cabinet outside the vibration chamber.

c. Connect the coaxial lead from the fitting at the rear of the CVTO to the CVTO INPUT (J12) at the rear of the main chassis.

d. Remove the CVTO power cord from its set of hooks at the rear of the main cabinet and plug it into the CVTO.

e. Remove the main power cord from its set of hooks at the rear of the main cabinet and connect it to an appropriate power source. An entry slot is provided for these cables under the cabinet door.

NOTE

It is important that the third wire conductor of the power cable be connected to a good ground.

NOTE

During operation, the reference oscillator and CVTO outputs and the frequency control voltage (above 30-mc operation) must be monitored. For this purpose a Hewlett Packard 410A (or 410B) vacuum-tube voltmeter is recommended. The meter probe is connected alternately to the oscillator test points. Above 30 mc, the d-c probe is connected to the frequency control test point (J17 at rear of chassis). A shielded extension cable may be used on the d-c probe, if necessary. The meter must be grounded to the unit at the a-c probe. In order to minimize probe capacity effects, the dummy vtvm probe supplied must be used at the open test point. This is particularly important above 50 mc. For ease of operation, two meters may be used.

f. Prepare the CVTO test block and crystal. Select the proper test block for the type of crystal to be tested. Refer to figure 32.

g. If the test crystal is in a type HC-6/U holder and is in the frequency range below 50 mc, use test block no. 4. Remove the cover with a coin, plug in the crystal, and replace the cover.

h. If the test crystal is in the frequency range above 50 mc and is in a type HC-6/U holder, use test block no. 1. Solder wires to the holder pins and insert the crystal into the block with wires through slot. Secure the crystal in the block with the screw-in

cover and solder the wire leads to the coaxial terminals of the block. Solder the compensating inductor across the terminal connections.

i. If the test crystal is mounted in a type HC-18/U holder, solder the crystal into test block no. 3. Use compensating inductor for frequencies above 50 mc.

j. If the test crystal is mounted in a type T-5-1/2 bulb (any frequency) solder it into test block no. 2. Use compensating inductor at frequencies above 50 mc. The volume between the sides of the test block and the glass bulb must be filled with potting compound. As suggested by the USASRDL, this is accomplished easily by pouring the cavity full of melted paraffin. If subsequent test characteristics indicate the need for another potting material, use a water-soluble compound such as artificial stone or dental casting investment. It usually is best to bake dry this potting material. For removal, the block may be soaked in water and the split mounting block removed. Further soaking will allow removal of the potting compound. In all cases, care should be taken to keep the wire leads from the crystal as short as possible and to keep the water-soluble compound from contacting these leads.

k. Mount the test block on the vibration table and connect the two small coaxial cables on the test block to the CVTO. Figure 33 provides information for selecting additional cable length as necessary between the crystal block and the CVTO. To use figure 33, proceed as follows: Find the working frequency along the lower edge of the chart. Follow the frequency line upward until it intersects one of the calibration bars. Follow the horizontal line to the left margin to determine the CVTO BAND setting and the proper length of cable to be added to that already on the test block. For bands V through VIII, note that the calibration bar extremes are marked with the logging scale limits.

NOTE

No precise frequency readout is intended. The chart of figure 33 is designed to provide cable length and BAND setting information. Final tuning must be done by oscilloscope indication.

Make sure the crystal cables are connected to the proper set of CVTO jacks for the frequency in use. Set the CVTO on a support which is isolated from the vibration section of the floor. Tape the coaxial cables to the shake table and the machine structure to minimize the length of cable being flexed.

4.4.1.2 Preliminary Adjustments.

a. With the function switch in STBY position, turn the equipment on. While the equipment is warming make the following settings:

b. Set the reference oscillator BAND control to the band which will include the desired frequency.

c. Set the reference oscillator TUNING control to the frequency desired.

d. Turn the reference oscillator CRYSTAL DRIVE control fully counterclockwise.

e. Plug the reference oscillator crystal (same frequency as that to be used for the CVTO) into the proper socket. Each set of sockets is composed of one socket for HC-6/U type crystals and one socket for HC-25/U type crystals. If the crystal in use (such as type HC-18) has wire leads, they must be soldered to the base of an HC-6/U holder. As indicated by the silk-screen markings, crystals for the range from 1 to 30 mc must be plugged into a socket in the left set and those for the range from 30 to 110 mc must be plugged into a socket in the right set. Above 50 mc, one of the furnished encapsulated inductors must be plugged into the companion socket to bring the crystal holder capacitance to antiresonance.

f. Set the phase detector BAND control to the band which will include the desired frequency.

g. Set the phase detector TUNING control to the desired frequency on the dial.

h. Set the CVTO BAND control according to information taken from figure 33.

i. Set the CVTO TUNING control according to information taken from figure 33. Figure 33 correlates the length of transmission line, band numbers, and the approximate dial setting necessary for any particular frequency between 1 and 110 mc. For bands I

through IV, consult figure 33 and use the horizontal line which gives the greatest frequency range above and below the desired frequency. In case of several choices, use the shortest length of line. On bands V through VIII, the CVTO dial is marked with a 0 to 100 linear logging scale. Figure 33 shows logging-scale numbers as a guide to the preliminary setting for tuning. Project the horizontal line to the left scale and read the length of line and the band setting to be used.

4.4.1.2.1 Reference Oscillator Adjustment.

a. The equipment has been warming. Set the function swich to R. O. ADJ position.

b. Turn the CRYSTAL DRIVE control clockwise until an output of 0.5 to 0.6 volt is measured at the R. O. TEST POINT.

c. Tune the reference oscillator until a peak output is obtained. If the output exceeds 0.7 volt, reduce the CRYSTAL DRIVE control setting until the output falls between 0.5 and 0.6 volt, and continue the tuning adjustment.

d. As a last step, reduce the crystal drive until an output of 0.5 to 0.6 volt is obtained at the peak output.

NOTE

The crystal dissipation is below military specifications limit with the oscillator producing the output of step d.

4.4.1.2.2 Crystal Vibration Test Oscillator Adjustment.

a. Set the function switch to CVTO ADJ postion.

b. Increase the setting of the CRYSTAL DRIVE LEVEL control until an output of 0.5 to 0.6 volt is measured at the CVTO TEST POINT.

c. Tune the CVTO over the complete dial. This is to become familiar with the band and length of transmission line in use. Oscillations other than that of the desired crystal frequency may occur. This is due largely to the addition of the transmission line to the oscillator circuitry and occurs only at the higher frequencies. During this step, the output should be kept below 0.7 volt.

d. Tune the oscillator to the crystal oscillation frequency. Peak up this output as in step c of paragraph 4.4.1.2.1.

e. Reset the CRYSTAL DRIVE LEVEL control so the reference oscillator and CVTO outputs are nearly equal. This condition should be maintained during each of the following steps.

NOTE

Crystal dissipation is well below military specified limit with the oscillator producing an output of 0.5 volt at all frequencies and combinations of transmission line so long as no equipment other than a HP-410B vtvm is connected to the CVTO TEST POINT. In nearly all cases there is ample margin for increased output voltage.

4.4.1.3 Final Adjustments.

4.4.1.3.1 Final Adjustments at Frequencies Below 30 Mc.

a. Set the function switch to CAL position.

b. Set the oscilloscope sweep and gain for easy viewing of the phase detector output.

c. Adjust the phase detector TUNING control for maximum output.

d. Retune the reference oscillator and the crystal vibration test oscillator until a beat note (seen on the oscilloscope) is about 4 to 9 cps/mc with the reference oscillator tuned to the lower side of the CVTO frequency. In most cases, this can be done by tuning only the reference oscillator.

e. Alternately adjust the phase detector TUNING and TRIM controls for the best triangular wave shape. This can usually be done best by tuning the main control for maximum output and the TRIM control for the best waveform. Record the peak-to-peak voltage. If a Polaroid camera attachment for the oscilloscope is available, it is recommended that a photograph be made of this calibration. On the photograph, draw lines extending the waveform slope at the zero-crossing to intersect at two adjacent peaks of the indication. The peak-to-peak voltage indication is taken between the points of slope intersection. See figure 58.

f. Move the function switch to LOCK position. The beat note seen on the oscilloscope should decrease in frequency until it finally becomes a horizontal line,

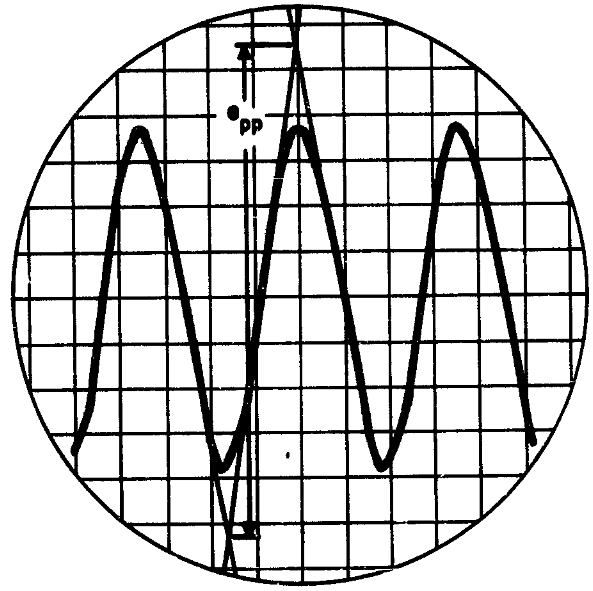


Figure 58. Calibration by Pr[illegible]ng Zero Slope

indicating that the reference oscillator has locked to the average CVTO frequency. If, instead, the beat note increases in frequency, the reference oscillator frequency is on the wrong side of the CVTO frequency. If the beat note decreases in frequency but does not lock, the beat note is too great and should be decreased. This can be done by retuning the reference oscillator slightly until lock is achieved.

g. Switch the oscilloscope vertical deflection from a-c to d-c. The oscilloscope trace should rise more than 3 volts but less than 1/4 of the actual peak-to-peak voltage recorded in step e. If this condition is not met, slight retuning of the reference oscillator will correct the situation.

h. Wait for five seconds after lock is achieved and then move the function switch to OPR position. The reference oscillator should remain locked to the CVTO signal. If it does not, the reference oscillator tuning should be retouched for lock condition and the oscillator output voltage rechecked. If this does not remedy the condition, recheck each previous step.

i. Return the function swtich to the CAL position and note whether or not the reference oscillator output voltage changes. The effect on the slope of the phase detector output by variation of oscillator output voltages can be checked by changing the crystal drive levels and noting the corresponding change in phase detector output. If the voltage varies enough to change the slope of the phase detector output, accurate calibration should be done after the crystal has been put through its test. Accurate calibration after the crystal test may be made by changing both crystal drive levels until the reference oscillator and CVTO outputs, when calibrated, are equal to those indicated during the test. Use the accurate calibration technique illustrated in figure 58. If the phase detector output change is insignificant with some small variation between oscillator output levels, the accurate calibration may be done before the crystal tests.

j. Return the function switch to LOCK position and wait five seconds until lock has occurred. Switch to OPR position. The equipment is now ready for the crystal vibration test.

NOTE

With the function switch in OPR position, the audio amplifier is in the circuit. This amplifies the phase detector output ten times and limits the measurement bandwidth to 4 kc. Note that the multiplication factor introduced must be considered in applying the calibration data.

k. The following alternate calibration procedure is suggested by USASRDL personnel. This procedure is described as a specific example.

Set up a beat note trace on the oscilloscope by adjusting the equipment as set forth in paragraph 4.4.1.3.1, through step e.

Set the oscilloscope sweep rate so that one-half cycle of the beat note covers six of the centimeter divisions (horizontally) on the oscilloscope grid. This represents 30 degrees per centimeter between the positive peak of the cycle and the negative peak (180 degrees).

Expand the oscilloscope time scale by a factor of 10. The presentation now represents three degrees per centimeter.

Adjust the vertical gain and position controls until the trace becomes the diagonal of a rectangle which is 2 cm (horizontal) by 6 cm (vertical). This produces a calibration of one degree per centimeter vertically for this gain setting. Make no further adjustments of the vertical variable sensitivity control on the oscilloscope. The system is now calibrated and all further adjustments of sensitivity will be in steps of 10. Keep gain multipliers in mind so that the actual calibration may be applied to the readout when the function switch is in OPERATE position.

Lock the system and start the shake table. Adjust either or both the fixed sweep rate or the fixed sensitivity to produce the desired presentation of the vibration information.

NOTE

Any convenient arrangement, beginning with 180-degree points of the trace spread horizontally across an integral number of centimeter grids, produces a calibration in a given number of degrees per centimeter.

4.4.1.3.2 Final Adjustments at Frequencies Above 30 Mc. Above 30 mc, a greater degree of care must be used in tuning the equipment. This results from the following:

The phase detector output decreases with increasing frequency and it becomes more likely that the d-c component will fall on a nonlinear portion of the unlocked phase detector output.

The reference oscillator and CVTO output voltages become more sensitive to load variations (phase detector tuning and standing waves on the interconnecting cables) as the frequency increases.

The circuit will hold-lock over a frequency range of approximately 30 cycles per second. This lock range is constant, making the equipment more difficult to adjust above 30 mc.

The principle used in locking the reference oscillator below 30 mc is based on increasing the frequency control voltage, requiring the reference oscillator frequency initially to be below the CVTO frequency. This principle is reversed for operation above 30 mc. In this range, the unlocked reference oscillator frequency should be above the CVTO frequency.

The range of beat notes which will produce a desirable lock condition is restricted. This restriction results from the specific difference between calibrate and lock frequency control voltages. Depending upon the unit in use, a beat note between 4 and 9 cps/mc is required.

a. Using the frequency control voltage, one can determine whether the beat note is too low or too high. In the CAL position, +1 volt d-c is applied to the Transcap circuit and to the integrator capacitor. In switching to LOCK position, this voltage will decay gradually. If lock condition first is produced and then broken as this voltage decays, it indicates the beat note is too low in frequency, producing feedback voltage insufficient to hold the lock. If, on the other hand, a lock condition is never attained, the beat note is too high. In either case the beat note should be changed accordingly. The reference oscillator may be retuned when the function switch is in the CAL or the LOCK position.

b. While the reference oscillator is changing frequency to lock condition, adjust the reference crystal drive in order to maintain a nearly constant output voltage. This need not be done continuously, but may be done in steps, allowing time for the circuit to stabilize between each change in drive level.

c. When lock condition is produced, the frequency control voltage should be within the range of -1 to +1 volt.

d. Switch the oscilloscope from a-c to d-c vertical input. The phase detector output is from 5 to 8 times the frequency control voltage. This ratio is maintained in OPR position and is used in determining whether or not the circuit is locked in the linear portion of the phase detector response. A frequency control voltage of ±1/4 volt is desirable.

For example, if the peak-to-peak phase detector output (seen in the CAL position) is 20 volts, it is best to operate with a d-c detector output of less than 5 volts. The ratio of d-c phase detector output to frequency control voltage (integrator output) is about 8. Therefore the frequency control voltage should be less than ±5/8 volt. The ±1/4 volt provides a safety factor, is attainable at all frequencies above 30 mc, and assures optimum calibration accuracy. If the frequency control voltage is other than ±1/4 volt, it may be adjusted by retouching the reference oscillator tuning.

e. Turn the function switch to OPR position. The lock may or may not hold. If the lock condition does hold, recheck the reference oscillator and CVTO outputs and the frequency control voltage. If they are within limits, the equipment is ready for the test. If the lock condition does not hold, check the reference oscillator and CVTO output voltages and retouch the reference oscillator tuning.

f. If, after several tries, a lock cannot be produced, the phase detector tuning should be rechecked to make sure the maximum output is available.

g. The equipment may be calibrated in two ways, depending upon the characteristics exhibited. If the reference oscillator and CVTO output voltages do not vary considerably from calibrate to operate conditions, calibration is best done before the tests are made. If, on the other hand, these output voltages do vary considerably, a rough calibration can be made before the test. After the test on a particular crystal, return the function switch to CAL position and reset both the reference oscillator and CVTO voltages to those values present during the test.

4.4.2 System Performance. The technical requirements for the equipment designed and constructed under this contract specify that the equipment must be capable of measuring a peak phase deviation of 3.13 millidegrees at 1 mc and 344 millidegrees at 110 mc when a crystal is being subjected to 10 g acceleration within the vibration range of 20 cps to 2000 cps. These data are presented in figure 59 by the specifications curve.

Also shown in figure 59 are a number of performance data points. These data were taken from a series of pictures such as are shown in figures 60, 61, 62, 63, 64, and 65.

These latter figures show the calibration curves, residual noise level, and phase deviation of crystals under vibration at several different signal frequencies. As can be observed, the resolution performance of the equipment is between 3 and 4 times better than required. At some operating frequencies, 60-cycle pickup is more pronounced than at others. In figure 60, it is interesting to note that the particular crystal under test shows a peak phase deviation under 5 g acceleration at 30 cps of 3.7 millidegrees. This is only slightly greater than the required resolution of the equipment with the crystal under 10 g acceleration.

Data taken on one 32-mc crystal previously supplied by USASRDL and tested on each of the final models resulted in the same phase deviation measurement to within ±5 percent.

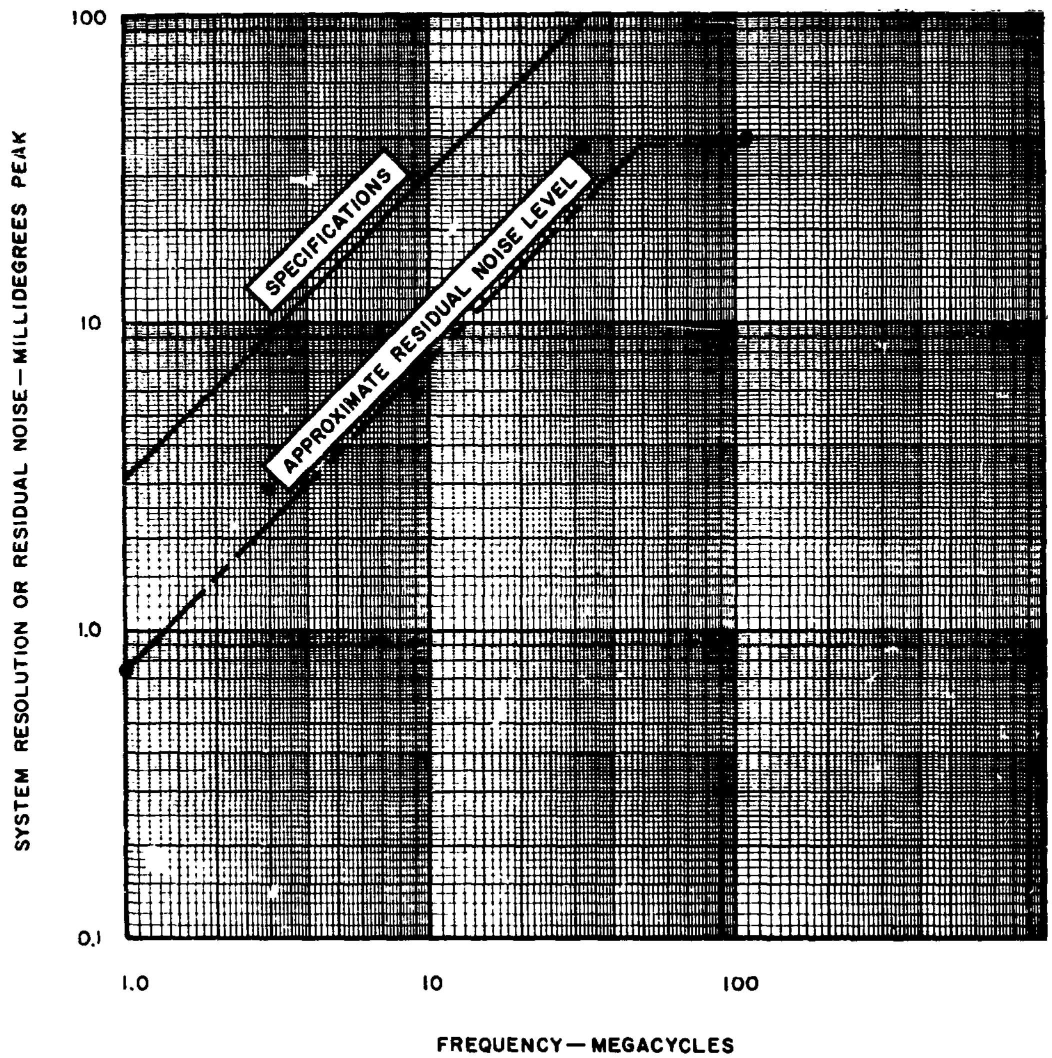
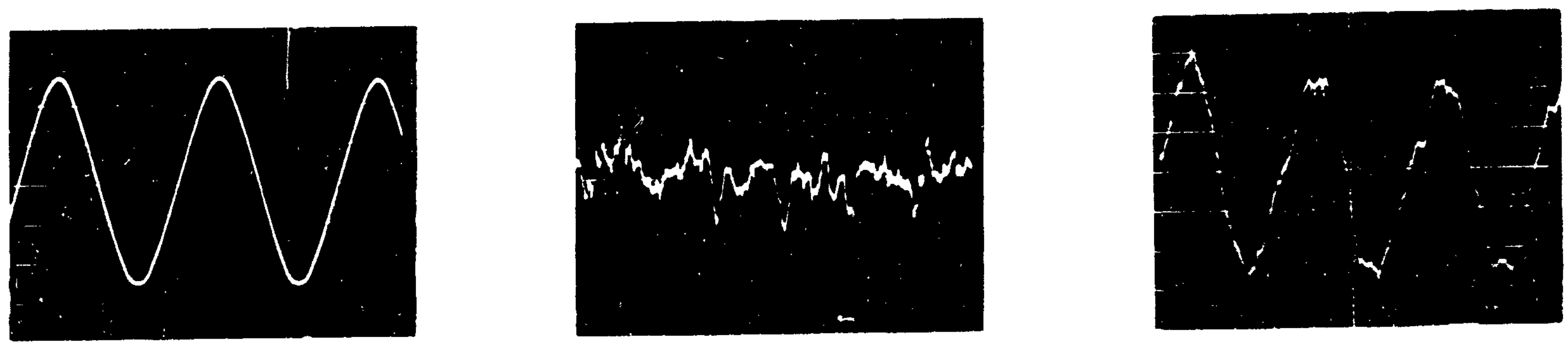


Figure 59. System Performance



CALIBRATION
SENSITIVITY : 30V/ DIVISION
SWEEP : 10 MS / DIVISION
SLOPE : 0.75 M°/ MV

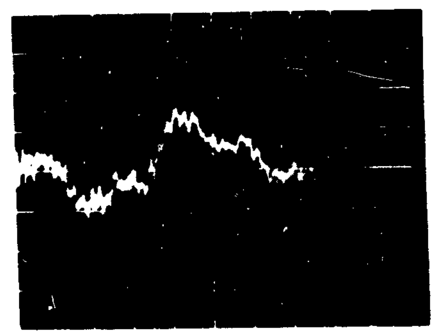
RESIDUAL NOISE
SENSITIVITY : 1 MV/ DIVISION
SWEEP : 10 MS / DIVISION
RESOLUTION : ≈ 0.75 M° PEAK

VIBRATION 5G AT 30 CPS
SENSITIVITY : 2 MV/ DIVISION
SWEEP : 10 MS / DIVISION
PHASE DEVIATION : 3.7 M° PEAK

Figure 60. System Performance at 1.0 Mc, Measurement System 101, USASRDL Crystal No. 7

CALIBRATION

SENSITIVITY : 30V/ DIVISION
SWEEP : 5 MS / DIVISION
SLOPE : 1.15 M°/ MV



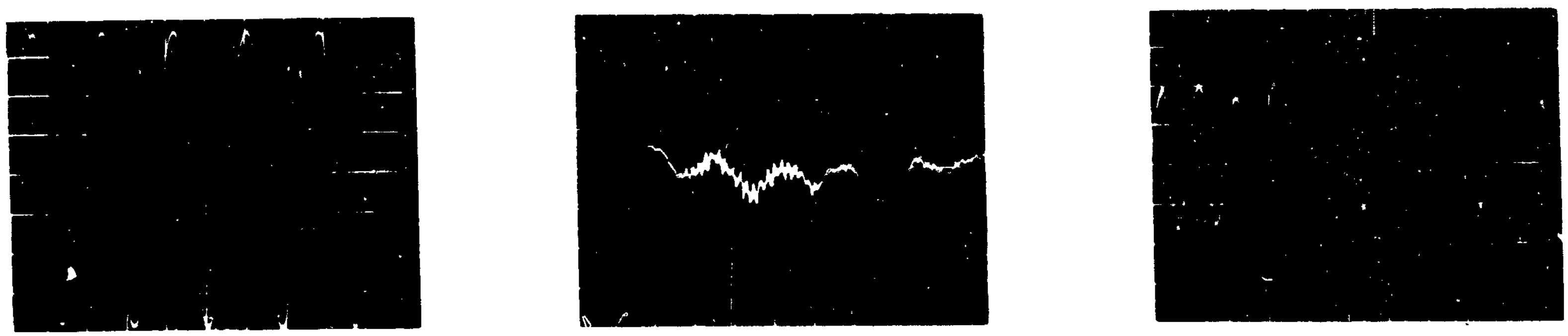
RESIDUAL NOISE

SENSITIVITY : 2 MV/ DIVISION
SWEEP : 10 MS / DIVISION
RESOLUTION : ≈ 2.9 M° PEAK

VIBRATION 10G AT 100CPS

SENSITIVITY : 30 MV / DIVISION
SWEEP : 5 MS / DIVISION
PHASE DEVIATION : 41 M° PEAK

Figure 61. System Performance at 3.0 Mc, Measurement System 101, McCoy Ruggedized Crystal



CALIBRATION

SENSITIVITY : 10V/ DIVISION
SWEEP : 2 MS / DIVISION
SLOPE : 1.5 M°/MV

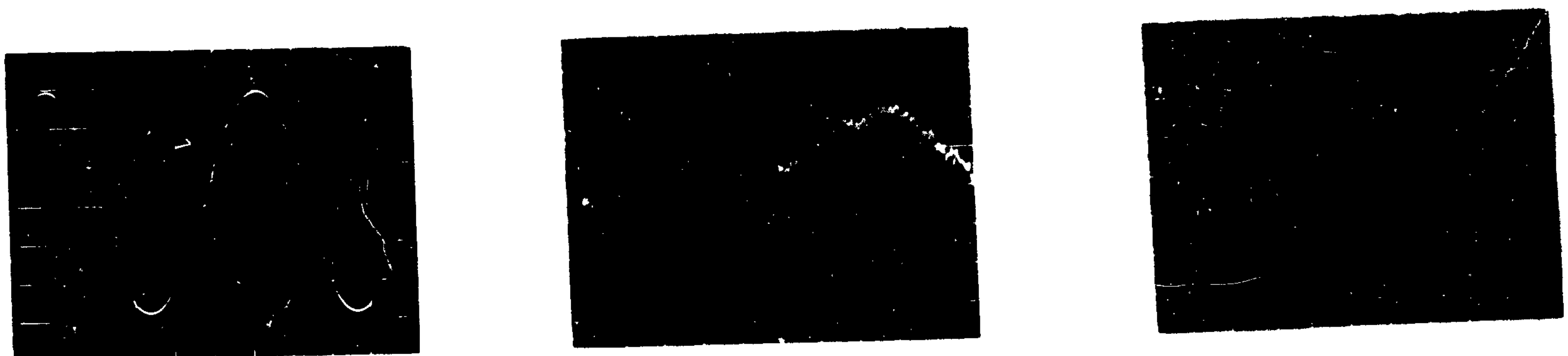
RESIDUAL NOISE

SENSITIVITY : 10 MV/ DIVISION
SWEEP : 10 MS / DIVISION
RESOLUTION : ≈ 6 M° PEAK

VIBRATION 10G AT 100CPS

SENSITIVITY : 20 MV/ DIVISION
SWEEP : 10 MS / DIVISION
PHASE DEVIATION : 23 M° PEAK

Figure 62. System Performance at 9.0 Mc, Measurement System 101, McCoy Ruggedized Crystal



CALIBRATION
SENSITIVITY : 10 V / DIVISION
SWEEP : 10 MS/DIVISION
SLOPE : 1.8 M°/MV

RESIDUAL NOISE
SENSITIVITY : 10 MV/DIVISION
SWEEP : 10 MS / DIVISION
RESOLUTION : 36 M° PEAK

VIBRATION 10G AT 100 CPS
SENSITIVITY : 20 MV/DIVISION
SWEEP : 10 MS / DIVISION
PHASE DEVIATION : 86 M° PEAK

Figure 63. System Performance at 32 Mc, Measurement System 101, USASRDL Crystal No. d4

CALIBRATION

SENSITIVITY : 10V / DIVISION
SWEEP : 2 MS/DIVISION
SLOPE : 2.25 M°/MV

RESIDUAL NOISE

SENSITIVITY : 10 MV/DIVISION
SWEEP : 10 MS /DIVISION
RESOLUTION : ≈ 8.6 M°PEAK

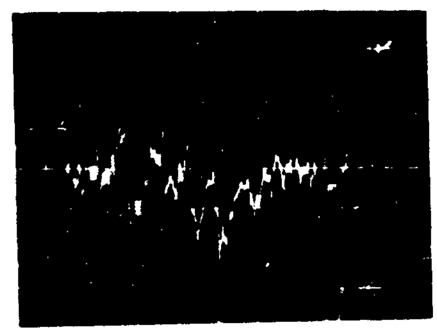
VIBRATION 10G AT 100 CPS

SENSITIVITY : 10 MV / DIVISION
SWEEP : 10 MS / DIVISION
PHASE DEVIATION : 83 M° PEAK

Figure 64. System Performance at 32 Mc, Measurement System 102, USASRDL Crystal No. d4

CALIBRATION

SENSITIVITY : 2 V / DIVISION
SWEEP : 0.5 MS / DIVISION
SLOPE : 8.6 M°/MV



RESIDUAL NOISE

SENSITIVITY : 3 MV / DIVISION
SWEEP : 10 MS / DIVISION
RESOLUTION : ≈ 39 M°

VIBRATION 5G AT 100 CPS

SENSITIVITY : 30 MV / DIVISION
SWEEP : 10 MS / DIVISION
PHASE DEVIATION : 700 M° PEAK

Figure 65. System Performance at 110 Mc Measurement System 101, McCoy Ruggedized Crystal

5. OVER-ALL CONCLUSIONS.

a. A tunable phase detector to cover the 1-mc to 110-mc frequency range is a practical device.

b. The system using the tunable phase detector, as constructed in the three final models, exceeded the measurement resolution specifications by a factor of at least three at all frequencies.

c. In order to build a successful tunable phase detector, it was necessary to provide compensating circuitry which amounted to three types of phase detector circuits in one assembly.

d. The tunable phase detector provided the simplest of the various equipment systems considered.

e. Measurement of the phase instability under vibration of the same crystal on each of the three final models resulted in agreement within 10 percent.

f. This measurement system provides an excellent analytical tool for the study of the phase stability not only of crystals but also oscillator circuits, frequency synthesizers, and transmitters.

g. Noise in various passive circuit elements also can be studied using this equipment.

h. Design equations for a phase-locked loop are available and very satisfactory.

i. The design and analysis of a wide-range crystal oscillator with the crystal operating in the middle of a coaxial transmission line is difficult.

j. The use of low-ripple d-c power supplies is mandatory for equipment of this extreme in resolution requirements.

k. The use of a pierce-oscillator circuit is practical for frequencies up to 110 mc. It is doubtful whether satisfactory performance can be obtained much higher in frequency with this circuit.

l. The sensitivity of equipment to acoustical vibrations can be reduced substantially by lining cabinet and chassis panels with thick styrofoam sheets.

6. RECOMMENDATIONS.

As a result of this contract, the following material is presented for consideration in improvement and/or further development of this, or similar, equipment:

6.1 Major Improvements.

a. The design and construction of a 10-kc or a 100-kc-to-1 mc multiplier-amplifier unit for use with signals lower in frequency than the range of the existing equipment.

b. The design and construction of a wide frequency range tunable amplifier for use in the study and development of low-level circuitry.

c. The design of a synchronous detector for use in conjunction with this equipment would make possible the detection of sideband levels at least 120 db below the carrier level.

d. Coupling between adjacent phase-detector-tuned circuits should be reduced. This can be accomplished by either installing shields between each section or shorting the unused transformers or tuned circuits.

e. A d-c isolation amplifier would be advantageous between either the output of the phase detector and the integrating network or the integrating network and the reference oscillator.

f. More isolation should be supplied between the reference oscillator and the crystal-vibration-test oscillator. This may be difficult to accomplish without increasing the residual noise level of the system.

g. Forced ventilation would be desirable but could cause an increase in microphonic problems if a blower were installed improperly.

h. If possible, construct the subassemblies for easier removal of the ceramic tubes.

i. Provision should be made to observe the d-c output of the phase detector when the function switch is in the OPR position.

j. If possible, because of their fragile nature, replace the 5- to 180-picofarad tubular glass capacitors in the phase detector with a 1- to 50-picofarad tubular glass capacitor, which is considerably more rugged. This will necessitate the use of more bands in the phase detector.

6.2 Minor Improvements.

a. A better voltage-variable reactance would be desirable for the vhf bands of the reference oscillator. One less sensitive to temperature would be particularly desirable.

b. Use steel instead of brass cams on the phase-detector tuning mechanism. The use of nylon guides on the lift rack should be considered.

c. The printed circuit board for the reference oscillator and CVTO should be changed to eliminate undesired or parasitic resonances in the ground structure.

d. Use Oldham couplers (phase-detector shaft coupling) with the center section fastened to one side.

e. Use a different type of coaxial connector in the equipment. Those used are too difficult to assemble but were the only ones obtainable within the required time.

f. Use a plug-in power connector with appropriate bypassing of leads on the reference oscillator.

g. Investigate the possible reduction of 60-cycle pickup by grounding the plus side of the heaters. Also investigate the possibility of raising the heaters approximately 50 volts above ground.

6.3 Further Development Work.

a. This equipment could be used to study the effect of noise in various components on circuit phase stability.

b. The development of a tunable phase detector and phase-locked system covering from 100 mc to 200 mc is entirely feasible for the study of crystal units and other applications in that frequency range.

7. IDENTIFICATION OF KEY PERSONNEL.

The following engineering personnel have engaged in various activities in conjunction with this project.

A. E. ANDERSON - Approximate hours 210.

From 1942 to 1945 Mr. Anderson served in the US Navy as an aviation electronics technician. He enrolled at the University of Minnesota after discharge from service, and received the BS degree in electrical engineering in March 1949. He joined Collins Radio Company after graduation and contributed to the design of uhf command control equipment and later served as head of the group responsible for the design of scatter communication equipment. Mr. Anderson is a member of Tau Beta Pi and Eta Kappa Nu, honorary engineering fraternities, and is a Senior Member of the IRE. He currently is assigned as Group Head of the Frequency Standards Group.

DR. D. E. NEWELL - Approximate hours 167.

Dr. Newell received the BS degree in electrical engineering from Iowa State University in 1952, and the MS degree in electrical engineering from the State University of Iowa in 1956. He worked as a graduate teaching assistant at the State University of Iowa from 1956 to 1958 and received his doctorate in electrical engineering in 1958. He joined Collins Radio Company in 1952 and worked on a part-time basis while attending graduate school. He was engaged in research and developmental projects concerned with frequency control, propagation, and communications. He conducted investigations into the characteristics of quartz crystals at liquid helium and liquid nitrogen temperatures. In September 1959 he joined the State University of Iowa as an assistant professor in the electrical engineering department.

Dr. Newell is a licensed professional engineer in the state of Iowa. He is a member of Eta Kappa Nu and Sigma Xi honorary fraternities. He also is a member of the Institute of Radio Engineers and the Iowa Academy of Sciences.

H. P. BROWER - Approximate hours 1820.

Mr. Brower was employed as a Radio Engineer by the Federal Bureau of Investigation, Washington, D. C., during the years 1942 through 1945. He returned to Oregon State College in 1946 and received a BSEE in 1948. From 1948 to 1950 he was employed by the Allis-Chalmers Manufacturing Company, Milwaukee, Wisconsin, first on their graduate training course and then as an assistant engineer in their Betatron Group. In 1950 he joined the staff of the University of California's Los Alamos Scientific Laboratory at Los Alamos, New Mexico, where he engaged in numerous aspects of atomic and hydrogen bomb development in that laboratory's Weapons Division. Mr. Brower began his work at Collins Radio Company in 1956 and was engaged in the Research and Development Division on a company-sponsored atomic frequency standard investigation. Since that time, he has served as Project Engineer on several frequency standard and frequency divider developments. He presently is Project Engineer on a USASRDL contract for the development of r-f phase-stability measurement equipment.

M. E. FRERKING - Approximate hours 1370.

Mr. Frerking enrolled in the University of Missouri in September 1954, and received a BS degree in electrical engineering in June 1958. At that time he joined Collins Radio Company and worked primarily with precision-crystal-test equipment. In July 1959, he entered the armed services and spent approximately one year at White Sands Missile Range, New Mexico. At White Sands, he was in the Engineering and Evaluation Section of the

Nike-Hercules missile project. During this period, he received approximately 20 percent of the credit hours required for a Masters Degree at New Mexico State University. In November 1960, he again rejoined Collins Radio Company and presently is working on r-f phase-stability measuring equipment.

DALE L. LINMAN - Approximate hours 680.

Mr. Linman received the BSME from Iowa State University in 1957. Upon completion of this training, he was employed at Interstate Power Company in Dubuque, Iowa, as an assistant to the Chief Production Engineer for a period of two years. Mr. Linman joined Collins in 1959, has been employed in the mechanical design of electronic equipment, and has worked in the areas of structural design, heat transfer, cam design, stress analysis, vibration, and machine dynamics.

More specifically, he has been engaged in design of the following equipment: 618V-1 and 618V-2 Mercury man-in-space transceivers, Autopositioner ® Repeater ARC-80, Variable Attenuator SRC-16 and the R-F Phase Stability Analyzer.

J. C. SCHMITT - Approximate hours 145.

Mr. Schmitt entered Speed Scientific School of the University of Louisville in 1956, and graduated with a Bachelor of Electrical Engineering in 1961. At that time, he joined the Collins Radio Company and began work on the R-F Phase Stability Analyzer.

Speed Scientific School employs a five year-cooperative work program. Under this program, Mr. Schmitt was employed for six months in the Electronic Laboratory of the Crane US Naval Ammunition Depot and three months each at the American Synthetic Rubber Company and Southern Bell Telephone and Telegraph Company.

While a senior, he presented a student paper on "A Generalized Tunnel Diode Amplifier/ Oscillator" at the region 5 AIEE Convention.

C. LEROY WILLEY - Approximate hours 147.

Mr. Willey received the BSME from Iowa State University in 1959. While attending school, he worked part-time at Deere and Company in Waterloo, as a student and junior engineer. He joined Collins Radio Company in 1959. Since that time, he has been engaged in the mechanical design of electronic equipment, and has worked in the areas of structural design, heat transfer, cam design, stress analysis, vibration, and machine dynamics. He also has worked in the specific area of accurate temperature control, as applied to crystal ovens.

More specifically, he has been engaged in design of the following equipment: R-F Oscillator ARC-80, Signal Comparator SRC-16, Transmitter Gain Control ARC-80, R-F Oscillator 618T, and the R-F Phase Stability Analyzer.

MOREY R. ZUBER - Approximate hours 285.

Mr. Zuber received a BS degree in mechanical engineering from Iowa State University in 1959. While attending Iowa State, he worked as a cooperative trainee with Deere and Company at Des Moines, Iowa, gaining experience in all manufacturing, management, and design functions. He joined Collins Radio Company in 1959, working in the field of mechanical design. This experience has included work in stress analysis, vibration, shock, structural design, gear design, and machine dynamics. This work included the ARC-80 switching repeater; ARC-80 relay assembly, mode select module; ARC-80 control, coupler servo module; SRC-16 environmental investigation; and the R-F Phase Stability Analyzer.

Acknowledgment also is given to the work performed on this project by laboratory technicians William A. Schwandt, George Jilek, and Glen Zirbes and to draftsmen John McDonough, Norman Wright, and Deene Speiler.

Styling credit should be given to the Industrial Design Group, headed by Theodore Papajohn.

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APPENDIX A

SUMMARY OF MIL-C-3098B

CRYSTAL UNITS, QUARTZ

APPENDIX A

SUMMARY OF MIL-C-3098B, CRYSTAL UNITS, QUARTZ

The technical requirements for this project specify that the end equipment include a suitable vibration table mounting for either HC-6, HC-18, or T-5 1/2 glass bulb crystal holders. In addition, the crystals are to be operated in conformance with the military standard for quartz crystals, MIL-C-3098B. This specification defines the maximum crystal resistance as a function of frequency and the maximum power that may be dissipated in a particular type of crystal. This information is summarized graphically in figures A-1 and A-2.

Figure A-1 shows the maximum allowable crystal drive power for the various types of crystals, as a function of frequency. The CR-48, CR-51, and CR-53 types were not considered applicable for test by this apparatus because they contain a pressure mounting. Above 10 mc, the CR-23, CR-52, CR-55, and CR-56 were used as the restrictive case in the design of the oscillator circuits because of their 2-milliwatt maximum drive level. Below 10 mc, the 10-milliwatt drive level of the CR-18 and CR-19 was used to determine the allowable crystal current. These maximum currents are plotted in figure A-2, together with the maximum specified crystal resistance allowed at a particular frequency.

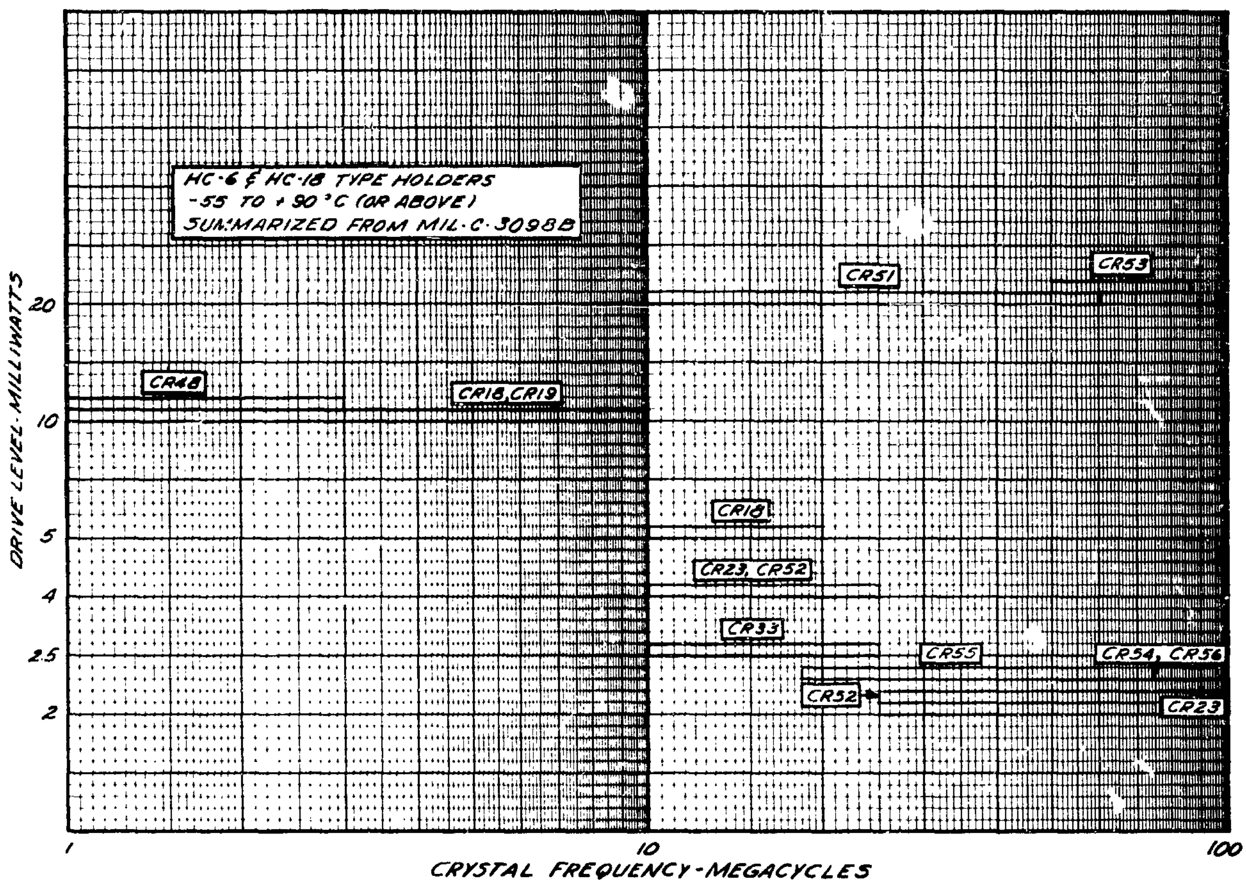


Figure A-1. Quartz Crystal Drive Level

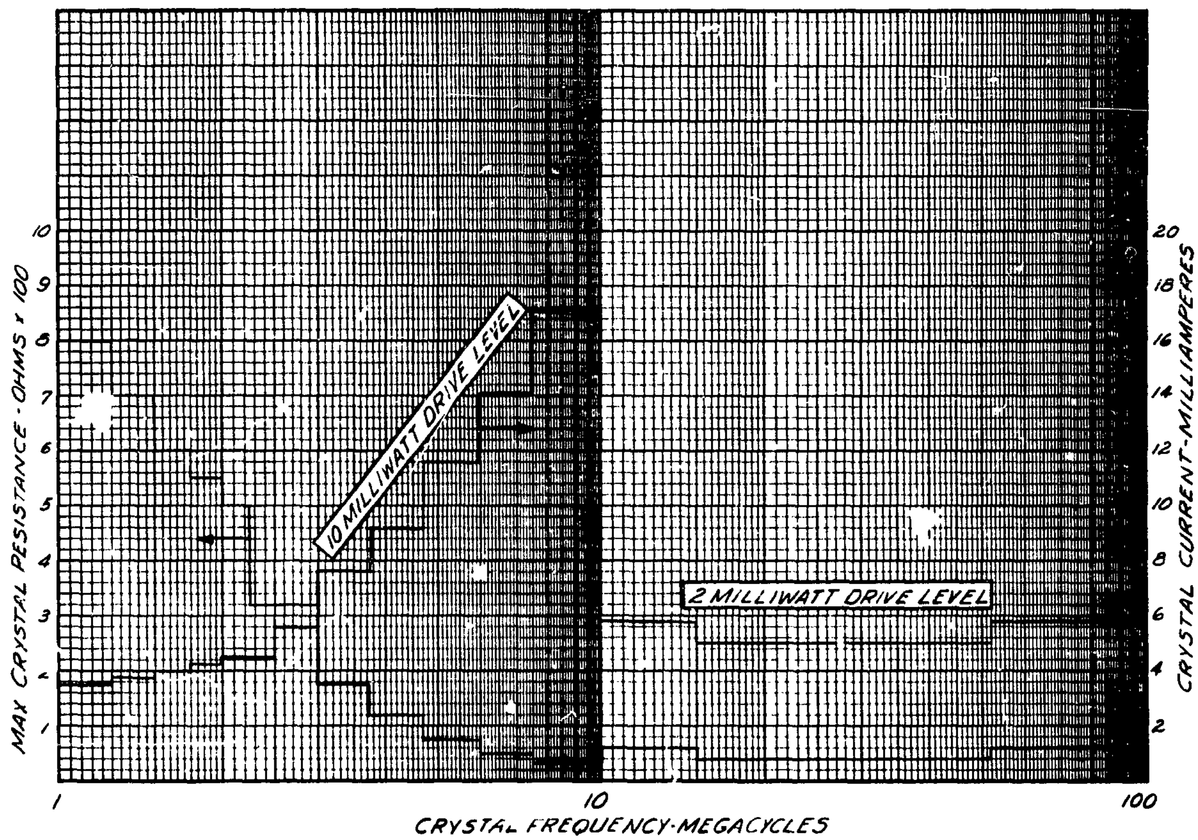


Figure A-2. Maximum Crystal Resistance and Crystal Current

APPENDIX B

COMPLETE SYSTEM SCHEMATIC DIAGRAM

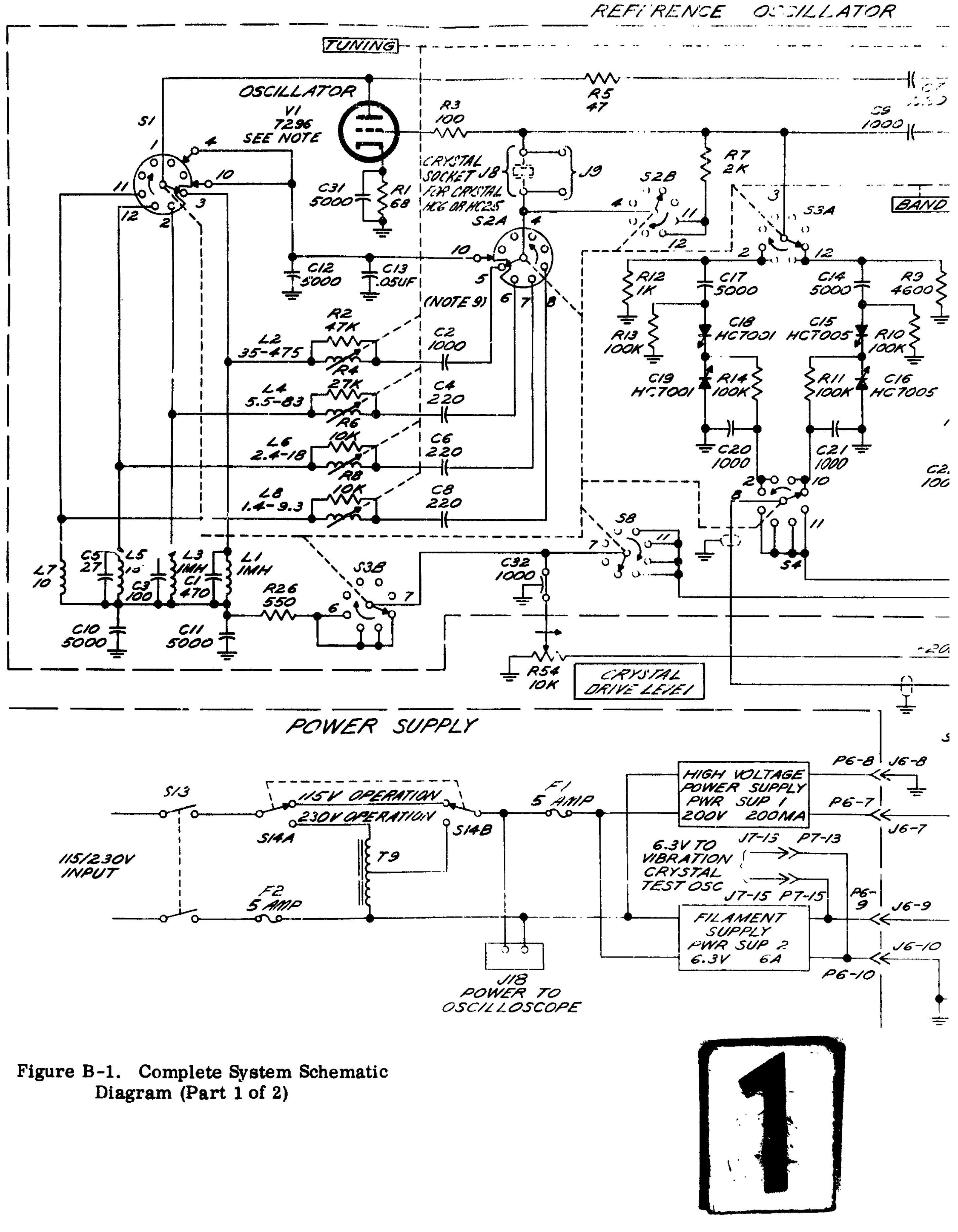
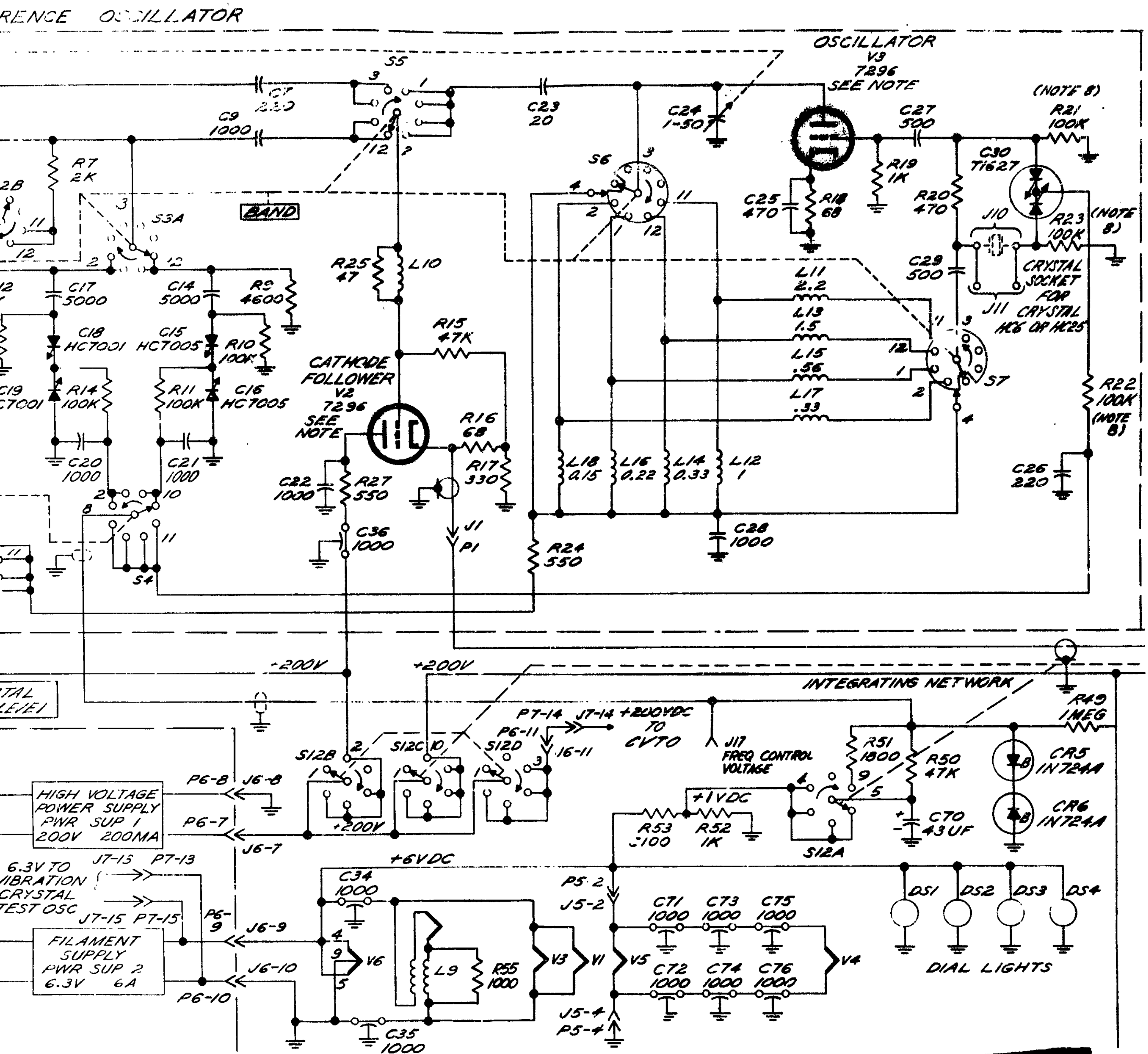
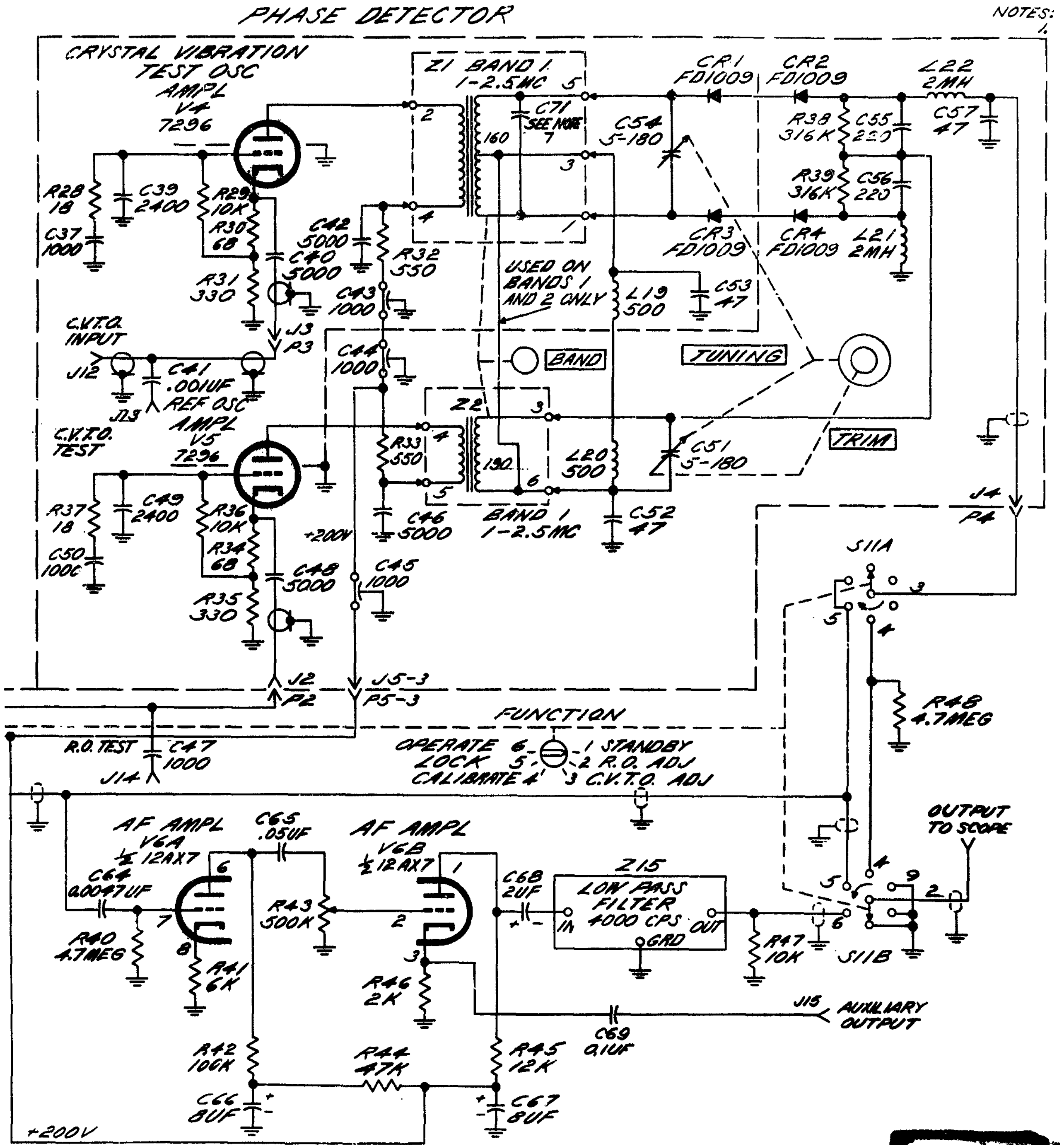


Figure B-1. Complete System Schematic Diagram (Part 1 of 2)

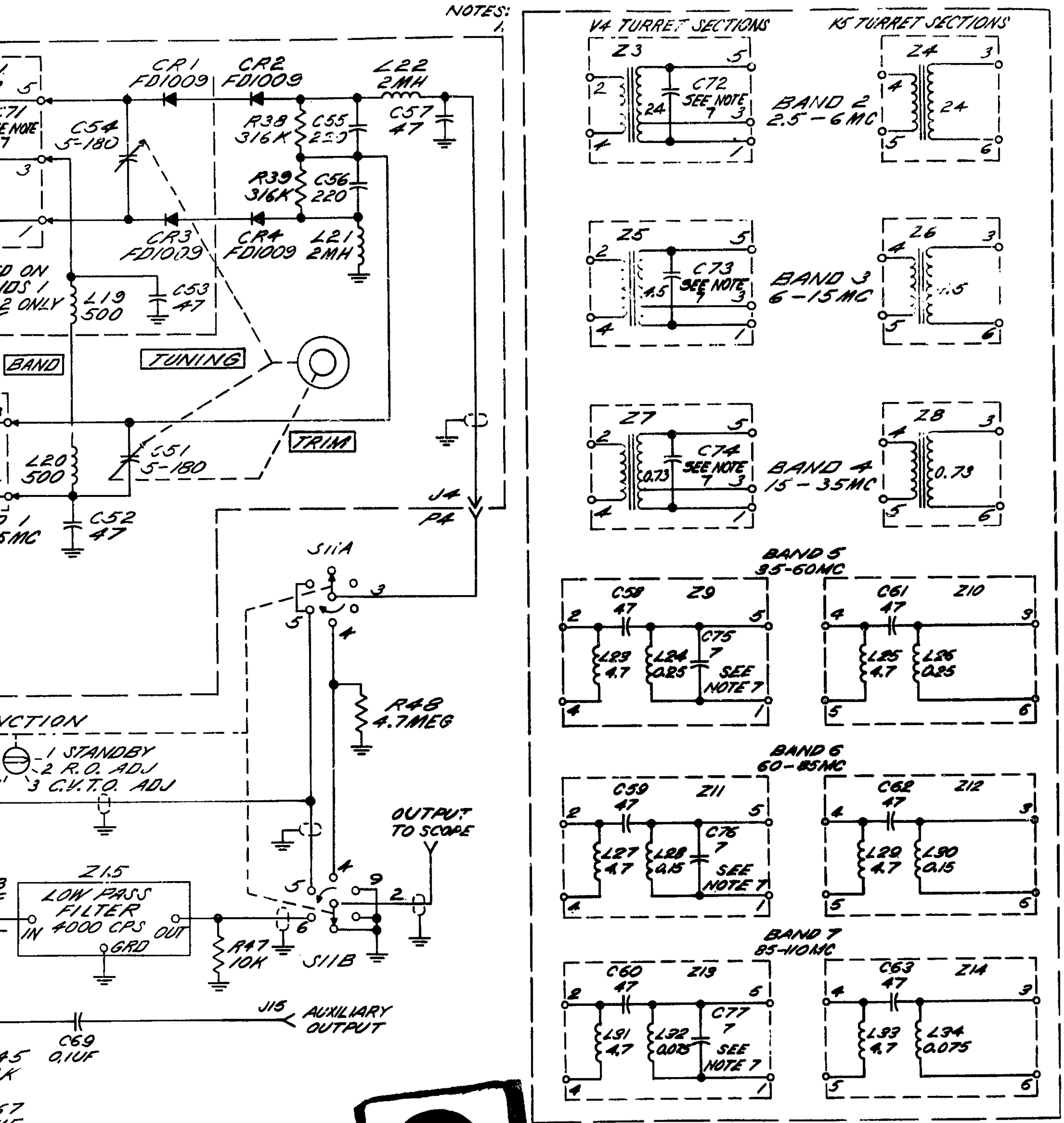

RENCE OSCILLATOR
OSCILLATOR V3 7296 SEE NOTE
S5
C7 220
C9 1000
C23 20
C24 1-50
C27 500
(NOTE 8)
R21 100K
C30 T1627
R7 2K
S6
BAND
S3A
R19 1K
C25 470
R18 68
R20 470
J10
R23 100K
(NOTE 8)
C29 500
CRYSTAL SOCKET FOR J11 CRYSTAL HC6 OR HC25
R25 47
L10
C17 5000
C14 5000
R9 4600
L11 2.2
L13 1.5
L15 .56
L17 .33
S7
C18 HC7001
C15 HC7005
R10 100K
R15 47K
CATHODE FOLLOWER V2 7296 SEE NOTE
R22 100K (NOTE 8)
C19 HC7001
R14 100K
R11 100K
C16 HC7005
R16 68
R17 330
C20 1000
C21 1000
L18 0.15
L16 0.22
L14 0.33
L12 1
C26 220
C22 1000
R27 550
C36 1000
J1
P1
C28 1000
R24 550
S4
-200V
+200V
INTEGRATING NETWORK
R49 1MEG
P7-14 J7-14 +200VDC TO CVTO
P6-11 J6-11
J17 FREQ CONTROL VOLTAGE
S12B
S12C
S12D
R51 1800
R50 47K
CR5 1N724A
CR6 1N724A
HIGH VOLTAGE POWER SUPPLY PWR SUP 1 200V 200MA
P6-8 J6-8
P6-7 J6-7
+200V
+1VDC
R53 100
R52 1K
S12A
C70 43UF
+6VDC
6.3V TO VIBRATION CRYSTAL TEST OSC
J7-13 P7-13
J7-15 P7-15
C34 1000
P5-2 J5-2
DS1 DS2 DS3 DS4
FILAMENT SUPPLY PWR SUP 2 6.3V 6A
P6-9 J6-9
J6-10 P6-10
V6
L9
R55 1000
V3 V1 V5
C71 1000 C73 1000 C75 1000
C72 1000 C74 1000 C76 1000
V4
DIAL LIGHTS
C35 1000
J5-4 P5-4

2



PHASE DETECTOR
NOTES:
1.
CRYSTAL VIBRATION TEST OSC AMPL
V4 7296
Z1 BAND 1 1-2.5MC
C71 SEE NOTE 1
160
CR1 FD1009
CR2 FD1009
L22 2MH
C57 47
C54 5-180
R38 316K
C55 220
R39 316K
C56 220
CR3 FD1009
CR4 FD1009
L21 2MH
R28 18
C39 2400
R29 10K
R30 68
C37 1000
C40 5000
C42 5000
R32 550
R31 330
C43 1000
USED ON BANDS 1 AND 2 ONLY
L19 500
C53 47
C.V.T.O. INPUT
J3
P3
J12
C41 .001UF
J13
REF OSC AMPL
V5 7296
C.V.T.O. TEST
C44 1000
BAND
TUNING
TRIM
Z2
R33 550
130
L20 500
C51 5-180
C46 5000
BAND 1 1-2.5MC
C52 47
J4
P4
R37 18
C49 2400
R36 10K
R34 68
C50 1000
C48 5000
+200V
C45 1000
R35 330
S11A
R48 4.7MEG
J2
P2
J5-3
P5-3
FUNCTION
OPERATE 6
LOCK 5
CALIBRATE 4
1 STANDBY
2 R.O. ADJ
3 C.V.T.O. ADJ
R.O. TEST
C47 1000
J14
AF AMPL V6A ½ 12AX7
C65 .05UF
AF AMPL V6B ½ 12AX7
C64 0.0047UF
R43 500K
C68 2UF
Z15 LOW PASS FILTER 4000 CPS
IN
OUT
GRD
OUTPUT TO SCOPE
R40 4.7MEG
R41 6K
R46 2K
R47 10K
S11B
J15 AUXILIARY OUTPUT
C69 0.1UF
R42 100K
R44 47K
R45 12K
C66 8UF
C67 8UF
+200V

1



2

2. V4 TURRET POS...
1 2 3
V5 TURRET POS...
1 2 3

3. TERMINAL VIEW OF ...
PLATE
GRID
CATHODE
FILAMENTS

4. REFERENCE OSC...
S1 T...

POSITION	
1	
2	
3	
4	
5	
6	
7	
8	

5. ALL SWITCHES ... ARROW INDICA... INCREASING PO... NUMBERS ON SW... TERMINAL DES... POSITION NUM...

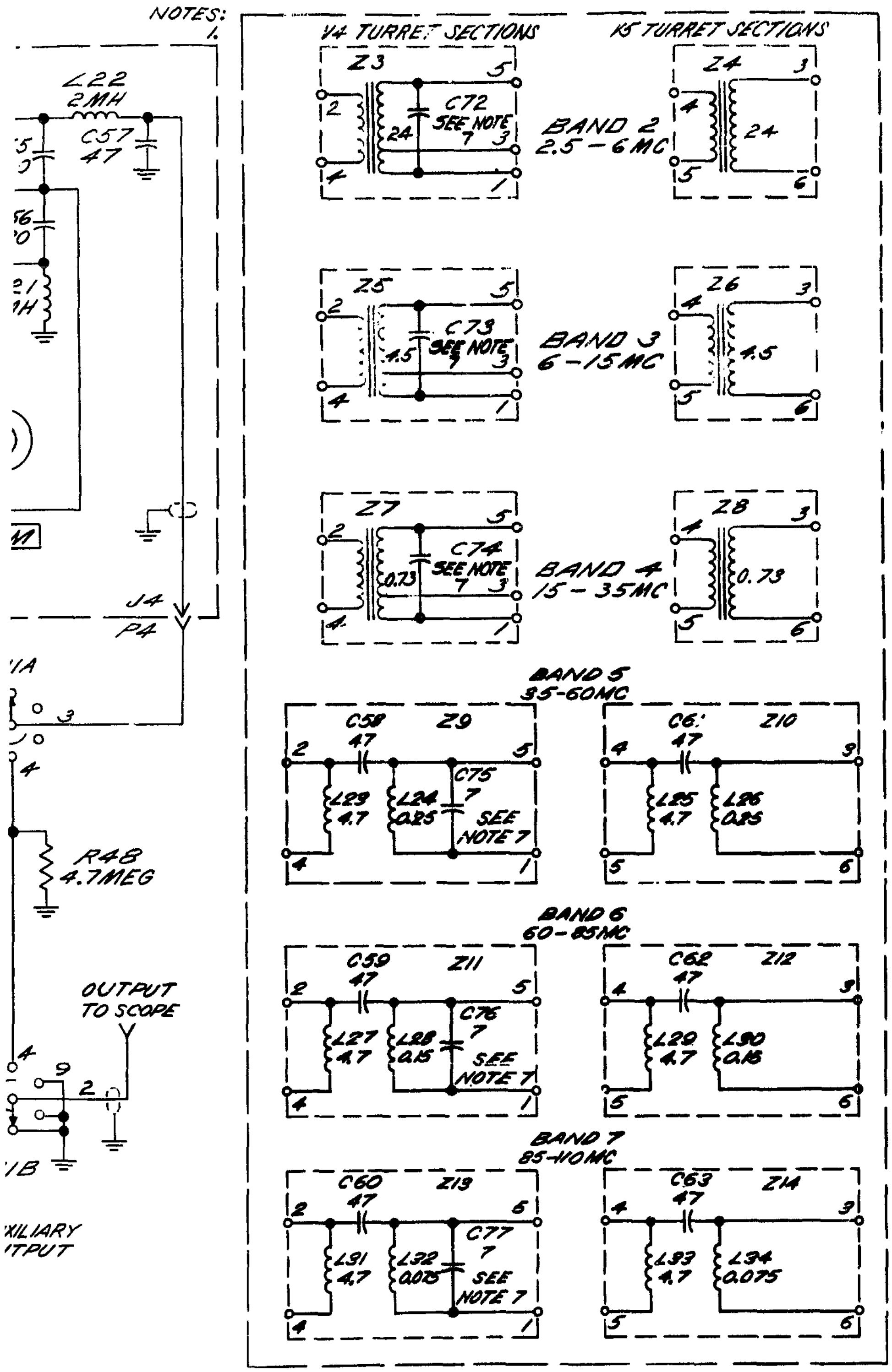
6. UNLESS OTHERW... RESISTANCE IS... INDUCTANCE ... AND ALL CA... PICOFARADS...

7. C71 THRU C77 SELE...

8. VALUES OF R21, R22, ... MANUFACTURE.

9. TERMINAL 6 OF S2A I... POSITIONS.

Figure B-1. C...
Diag...



2. V4 TURRET PORTION CONTACTS

1 2 3 4 5

V5 TURRET PORTION CONTACTS

1 2 3 4 5 6

3. TERMINAL VIEW OF 7296 TUBE PIN CONNECTIONS

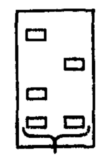


PLATE
GRID
CATHODE
FILAMENTS

4. REFERENCE OSCILLATOR BAND SWITCH
S1 THRU S8

POSITION	FREQUENCY
1	1 - 3 MC
2	3 - 10 MC
3	10 - 18 MC
4	18 - 30 MC
5	30 - 40 MC
6	40 - 60 MC
7	60 - 85 MC
8	85 - 110 MC

5. ALL SWITCHES SHOWN IN POSITION ONE. ARROW INDICATES DIRECTION OF INCREASING POSITION NUMBER. NUMBERS ON SWITCHES INDICATE TERMINAL DESIGNATION AND NOT POSITION NUMBER.

6. UNLESS OTHERWISE INDICATED ALL RESISTANCE IS IN OHMS, ALL INDUCTANCE IS IN MICROHENRYS AND ALL CAPACITANCE IS IN PICOFARADS.

7. C71 THRU C77 SELECTED IN MANUFACTURE.

8. VALUES OF R21, R22, AND R23 SELECTED IN MANUFACTURE.

9. TERMINAL 6 OF S2A IS UNGROUNDED IN ALL POSITIONS.

3

Figure B-1. Complete System Schematic Diagram (Part 2 of 2)

APPENDIX C

DELIVERY OF REPRODUCIBLE SHOP DRAWINGS

TO USASRDL

APPENDIX C

DELIVERY OF REPRODUCIBLE SHOP DRAWINGS TO USASRDL

One set of reproducible shop drawings has been delivered to USASRDL as part of this report. These drawings will enable a competent technician to construct a similar equipment.

APPENDIX D

FINAL REPORT DISTRIBUTION LIST

APPENDIX D

FINAL REPORT DISTRIBUTION LIST

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This contract is supervised by the Solid State and Frequency Control Division, Electronic Components Department, USASRDL, Fort Monmouth, New Jersey. For further technical information contact Mr. George H. Gougoulis , Project Engineer, Telephone: 535-1214.